



(12) **United States Patent**  
**Blom**

- |      |   |                     |      |         |                |                       |
|------|---|---------------------|------|---------|----------------|-----------------------|
| (54) | <b>CHOPPER STABILIZED AMPLIFIER WITH SYNCHRONOUS SWITCHED CAPACITOR NOISE FILTERING</b> | 2014/0210547        | A1 * | 7/2014  | Tomioka .....  | H03F 3/387<br>330/9   |
|      |   | 2015/0207477        | A1 * | 7/2015  | Stanescu ..... | H03F 3/45475<br>330/9 |
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*H03F 1/26* (2006.01)  
*H03F 3/393* (2006.01)

(52) **U.S. Cl.**  
 CPC ..... *H03F 1/26* (2013.01); *H03F 3/393*  
 (2013.01); *H03F 2200/171* (2013.01)

(58) **Field of Classification Search**  
 CPC ..... H03F 1/02  
 USPC ..... 330/9; 327/124, 307  
 See application file for complete search history.

(56) **References Cited**

## U.S. PATENT DOCUMENTS

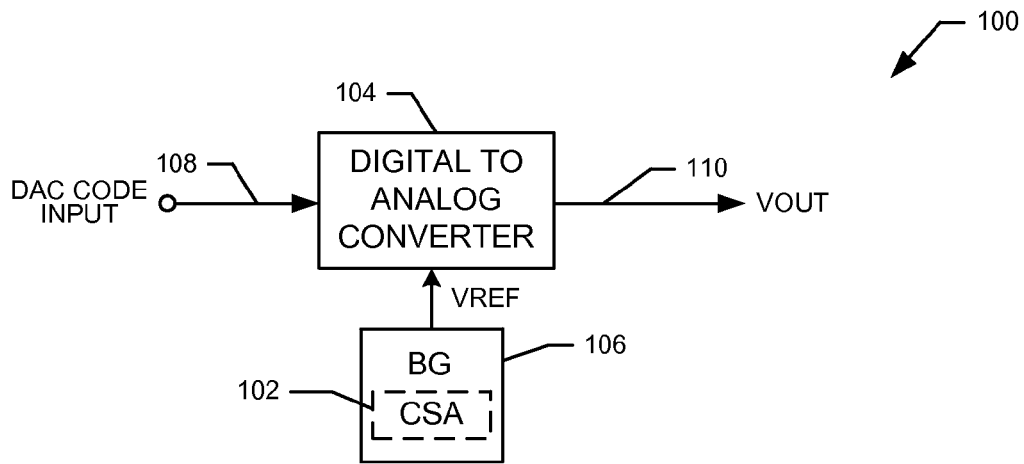


FIG. 1

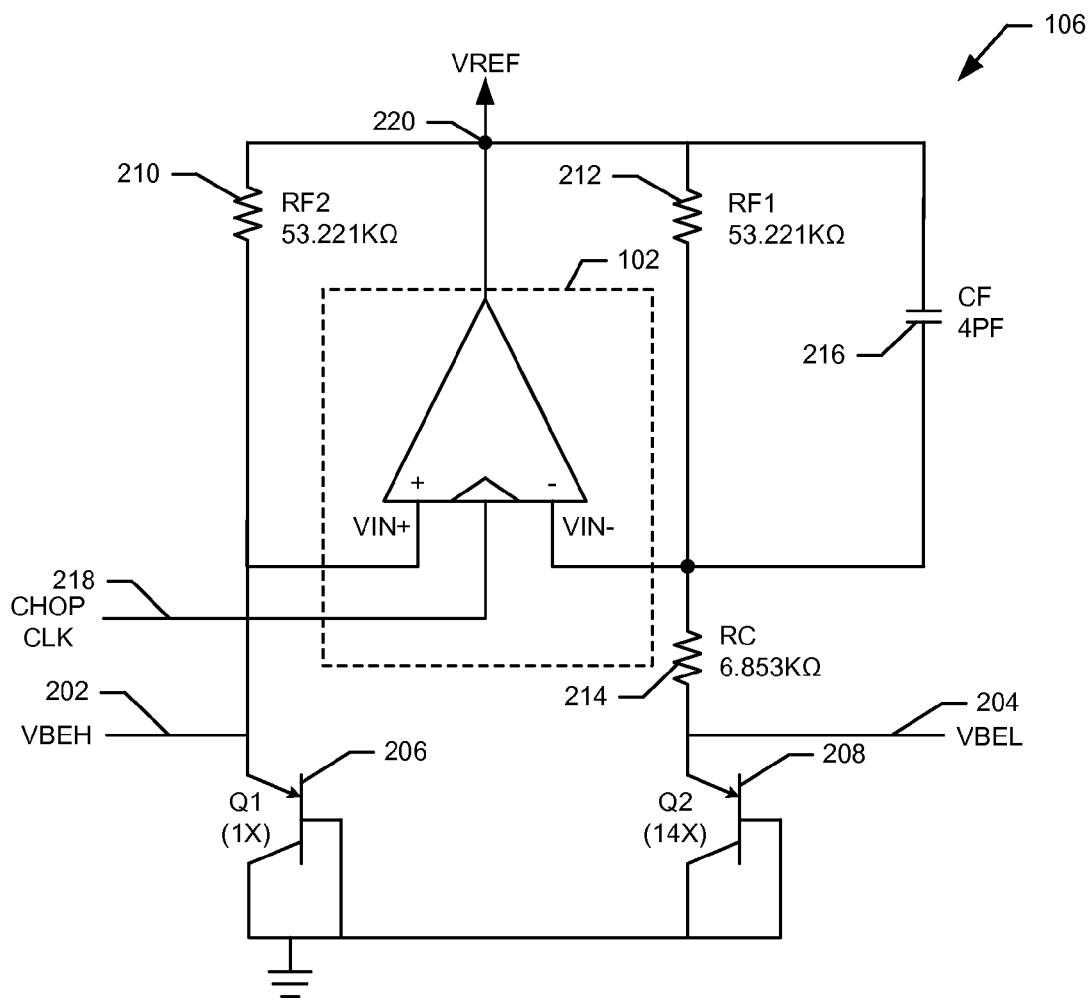


FIG. 2

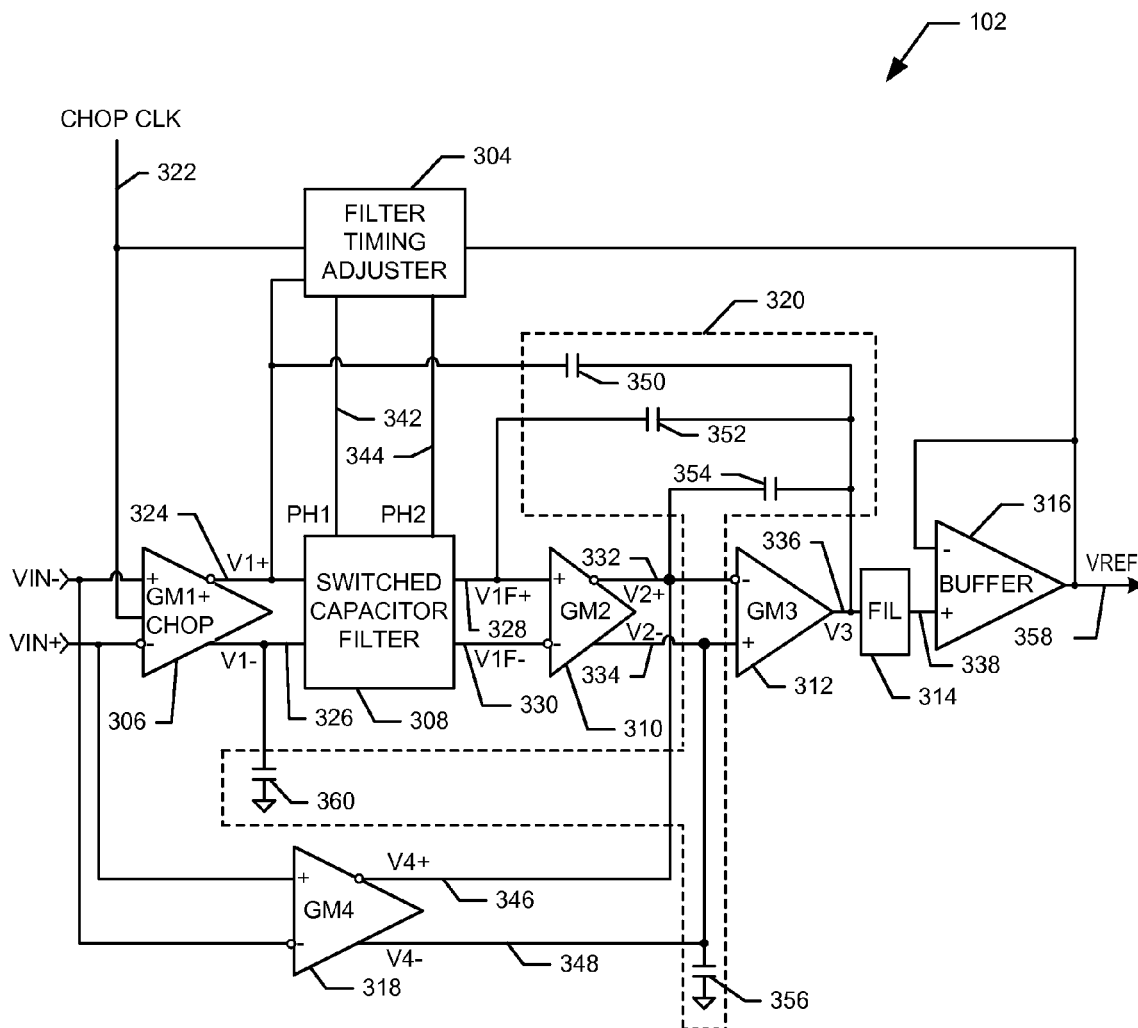


FIG. 3

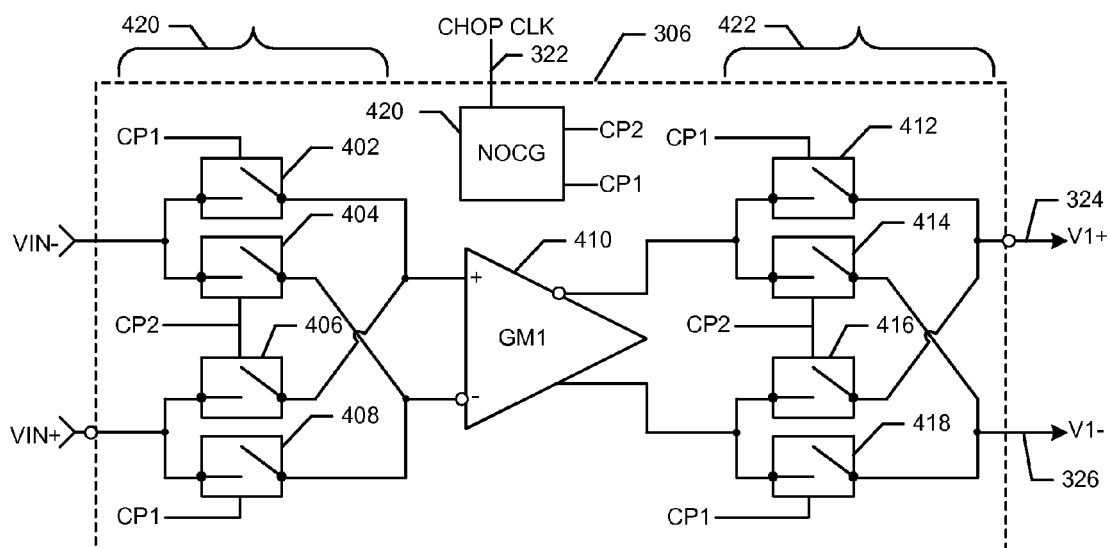


FIG. 4

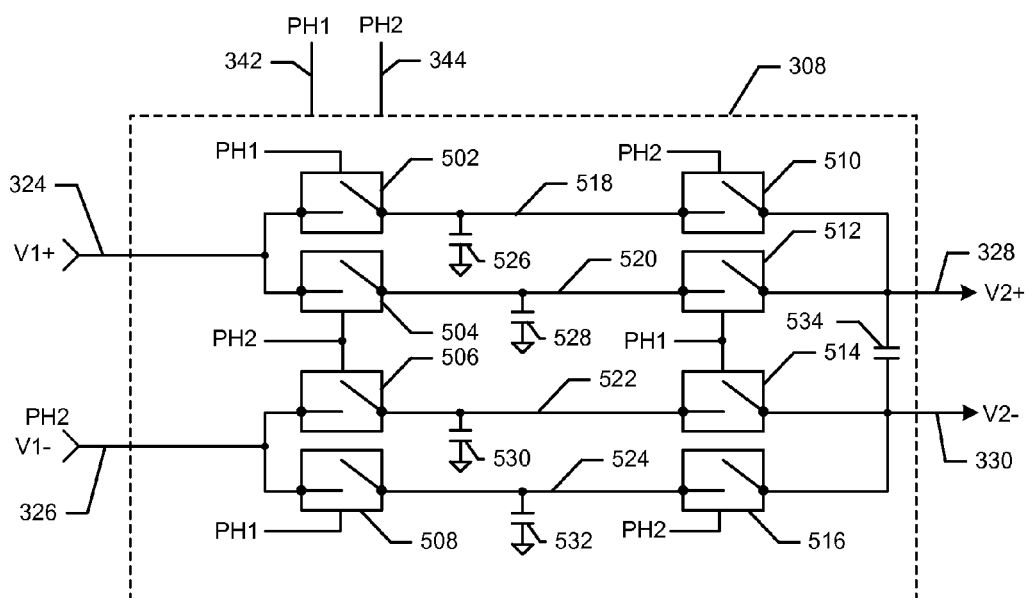


FIG. 5

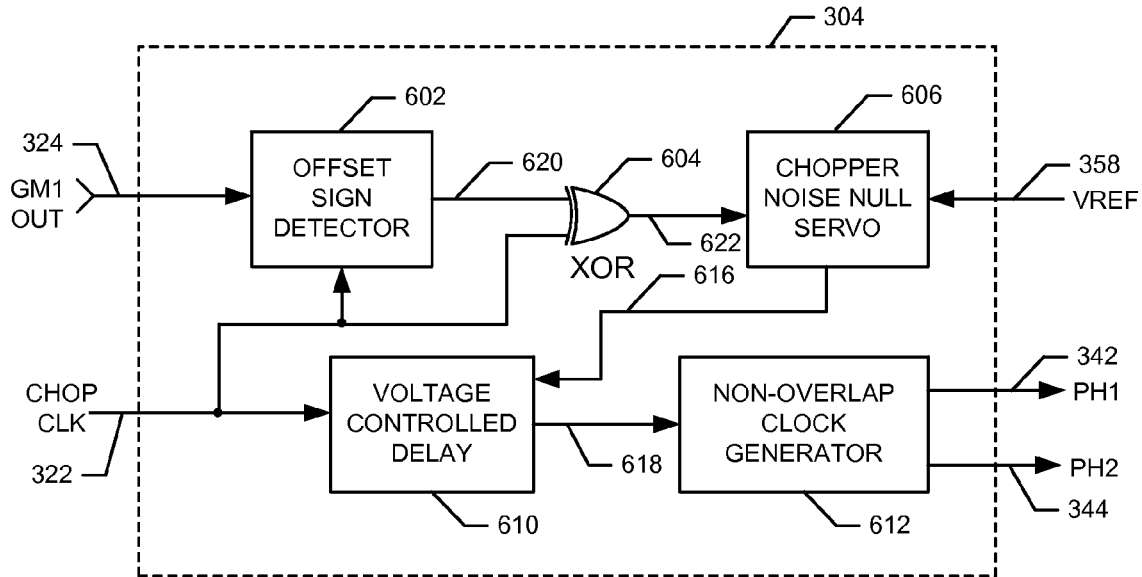


FIG. 6

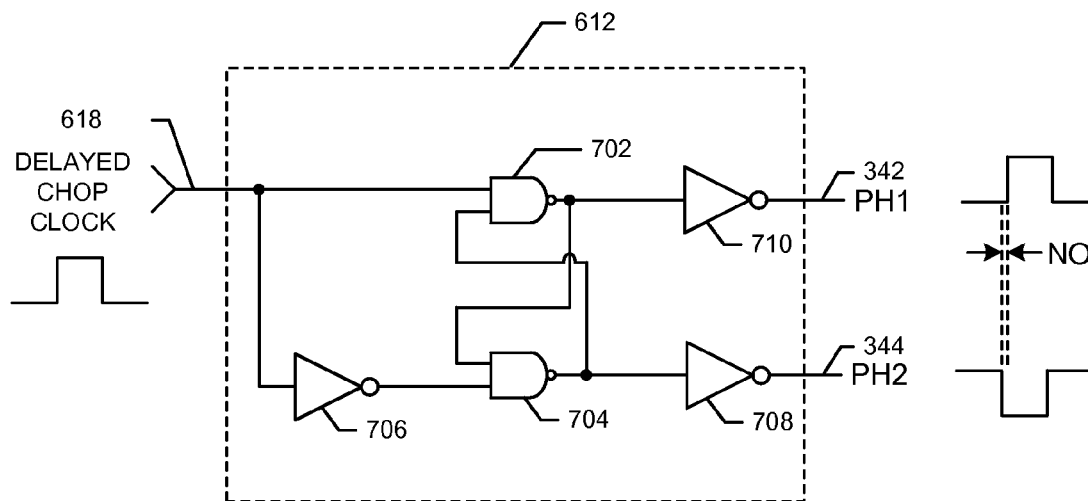


FIG. 7

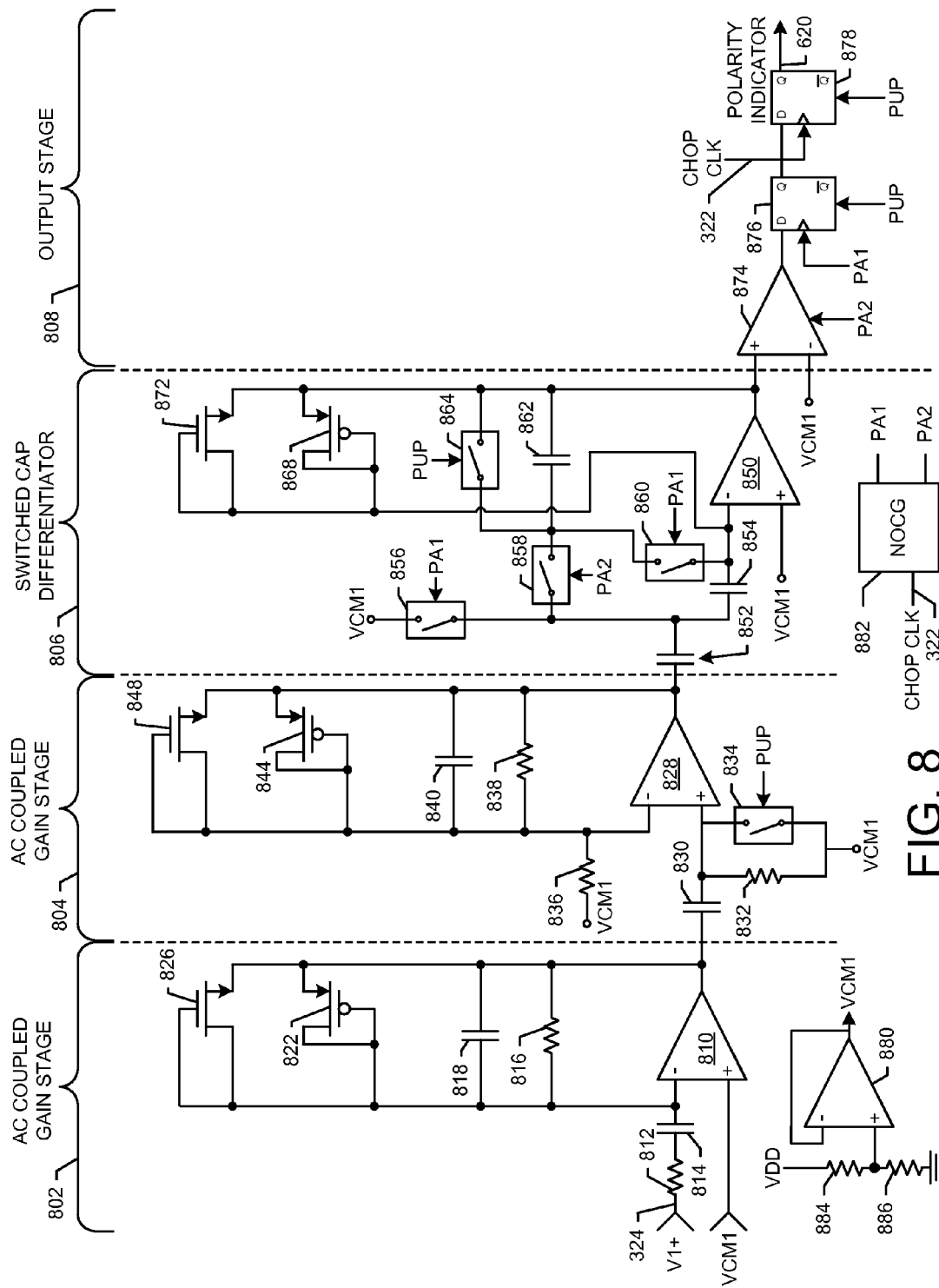


FIG. 8

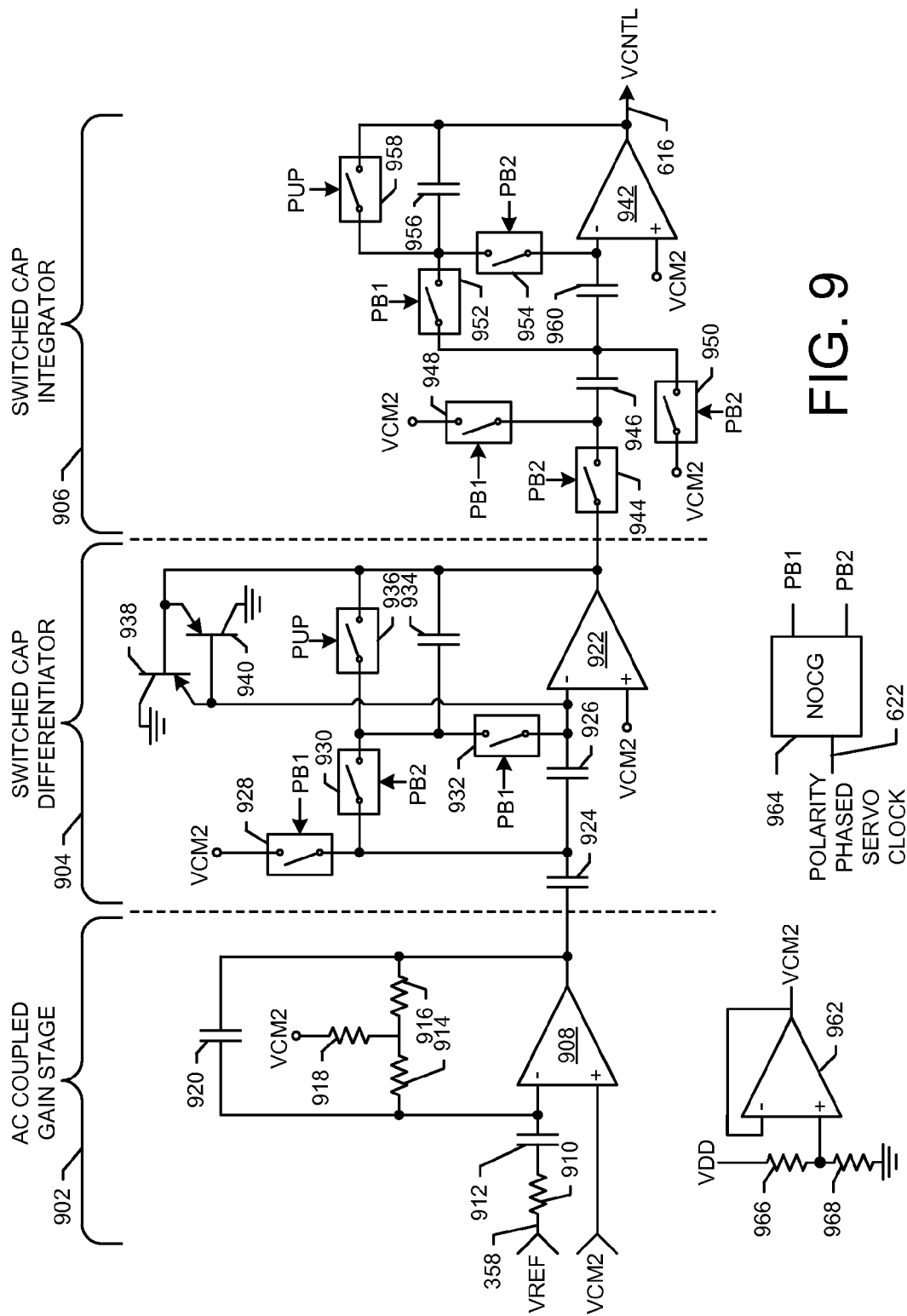


FIG. 9

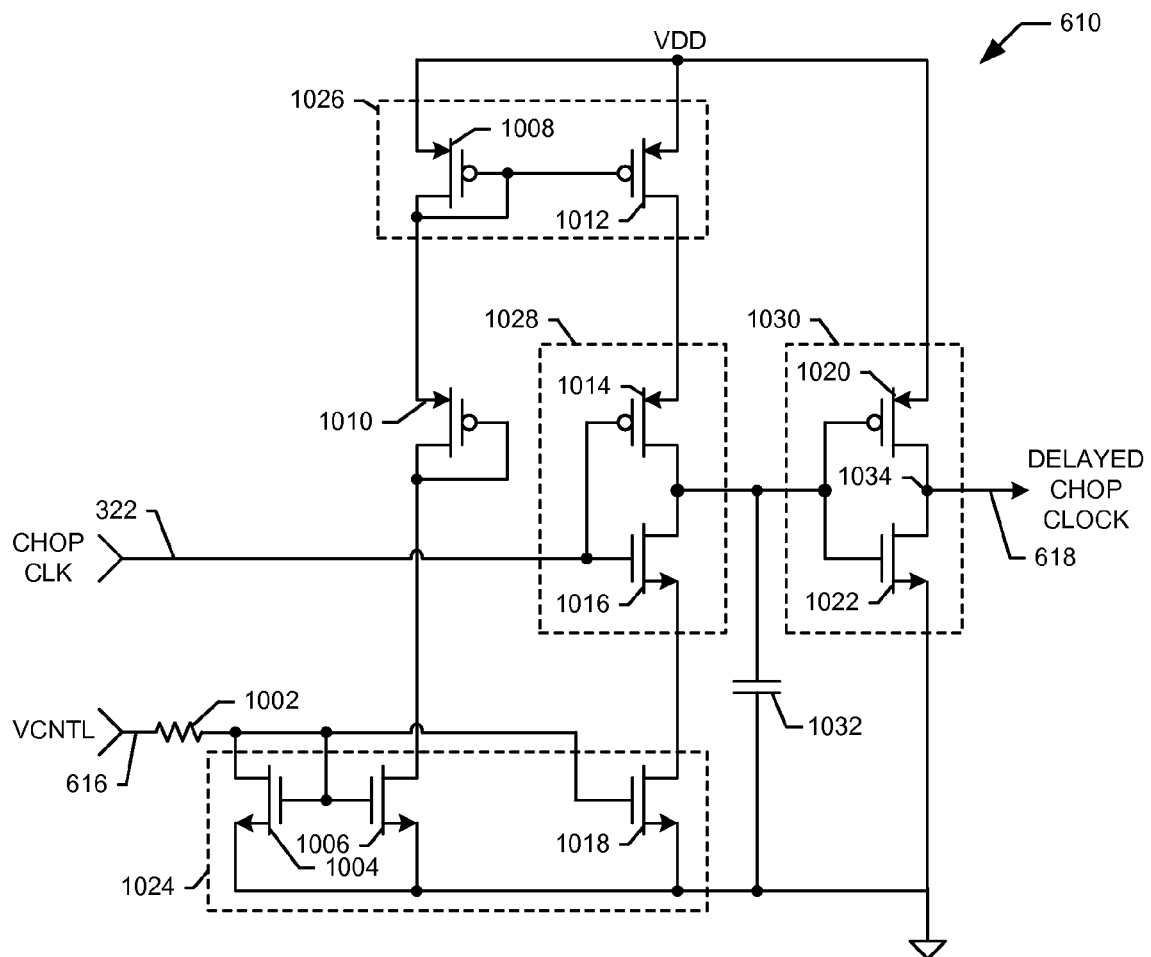


FIG. 10

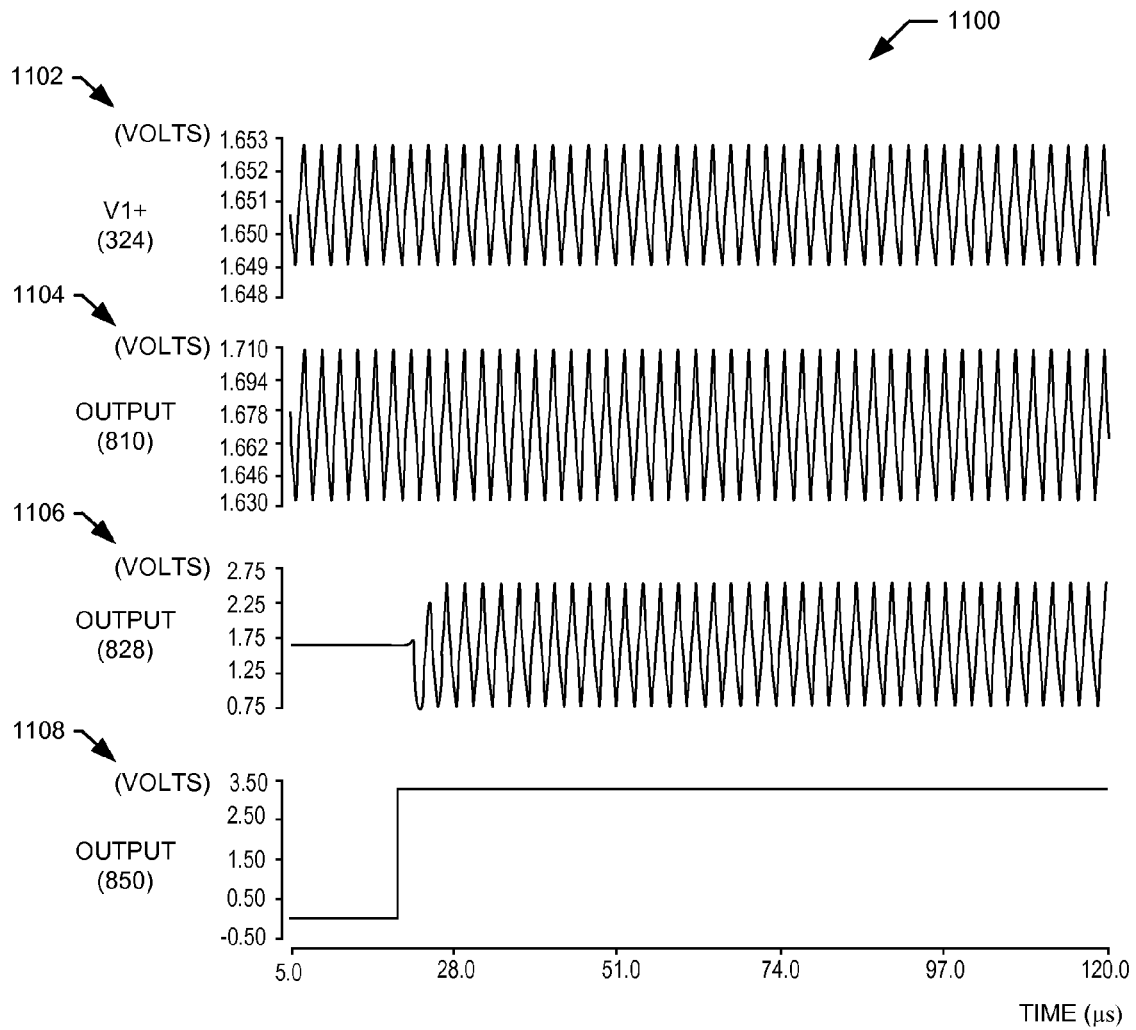


FIG. 11

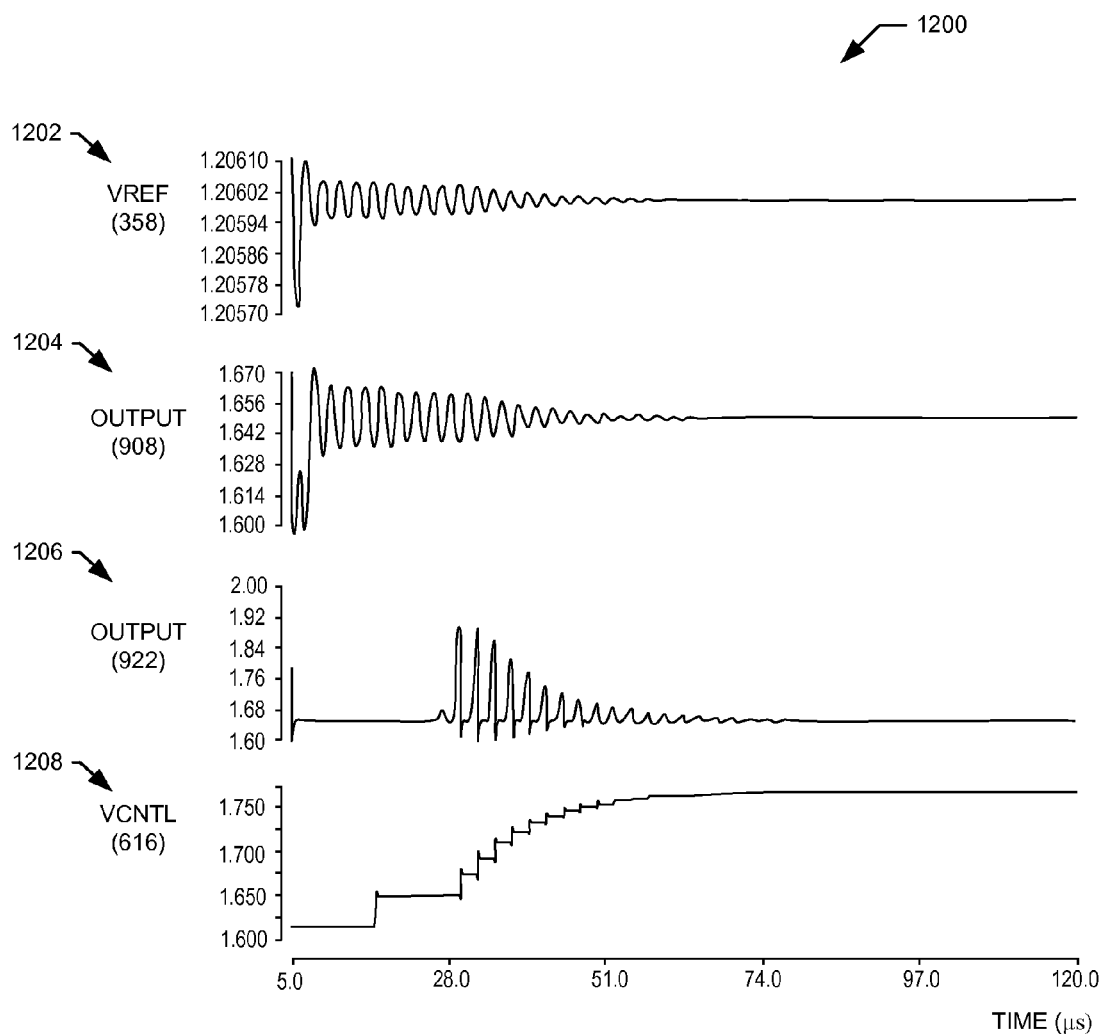


FIG. 12

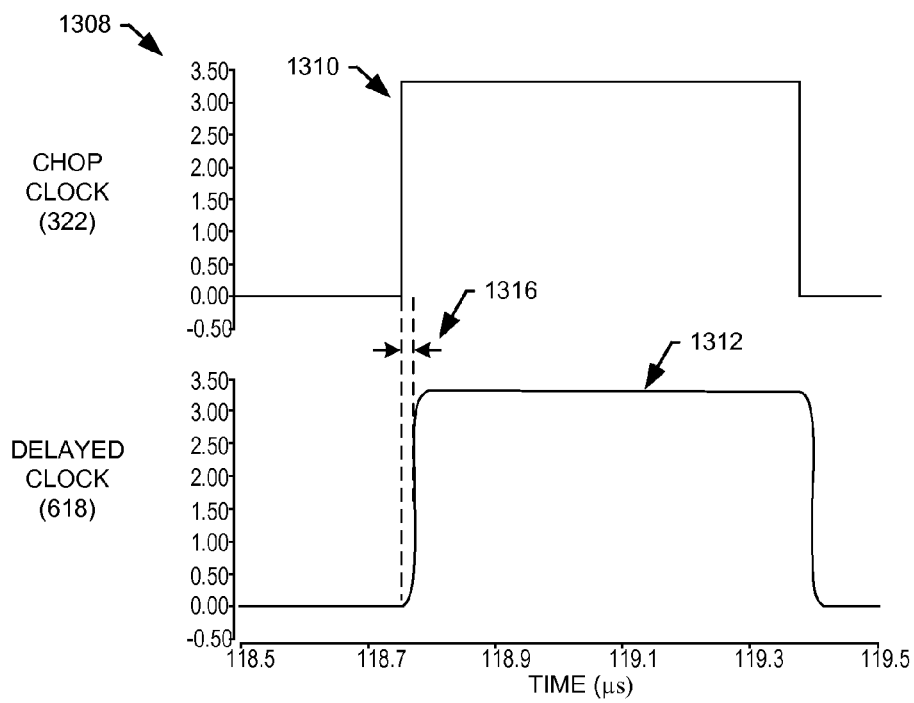
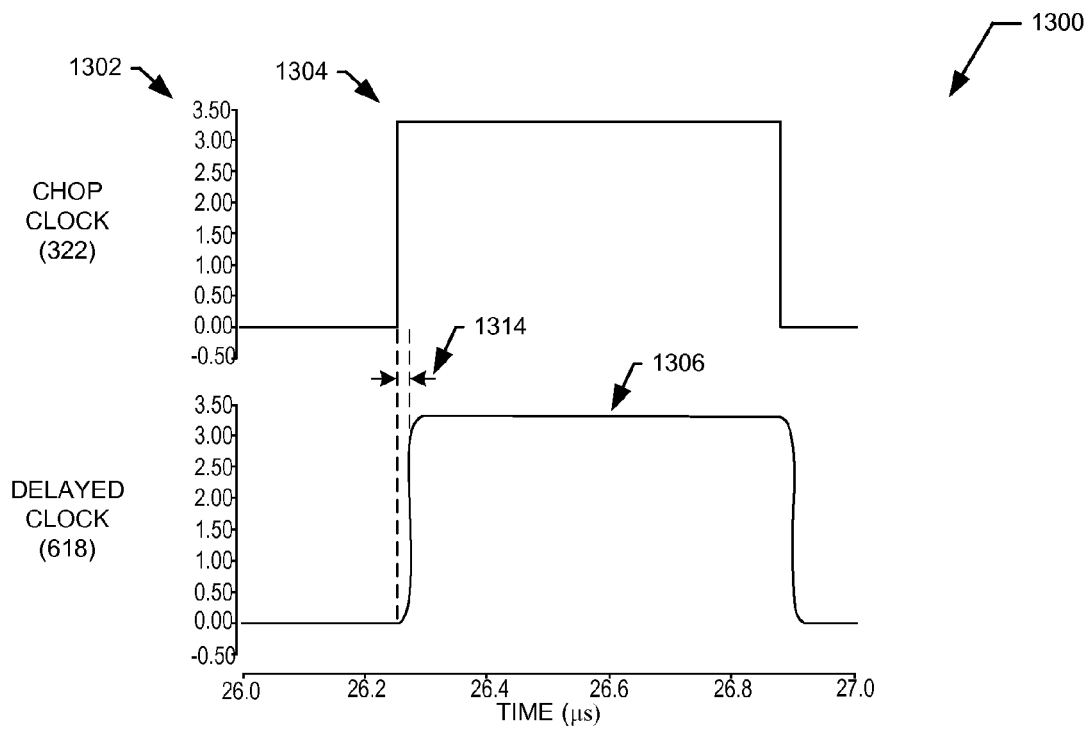


FIG. 13

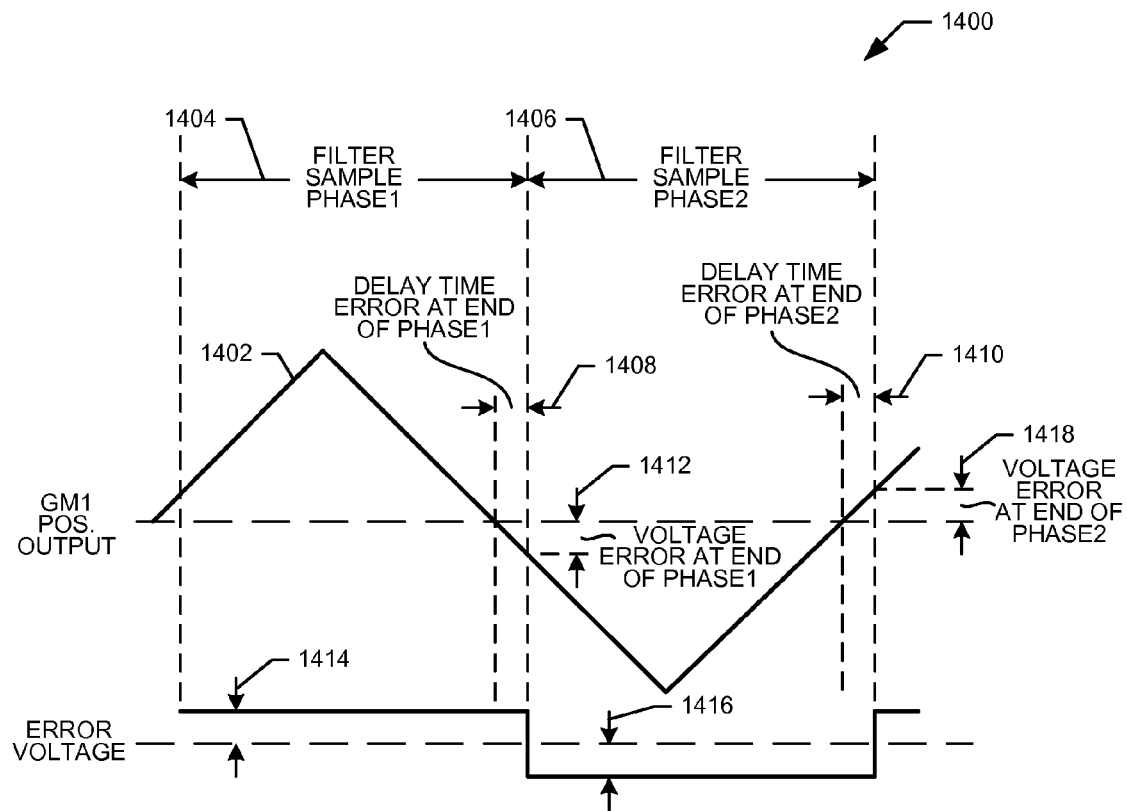


FIG. 14

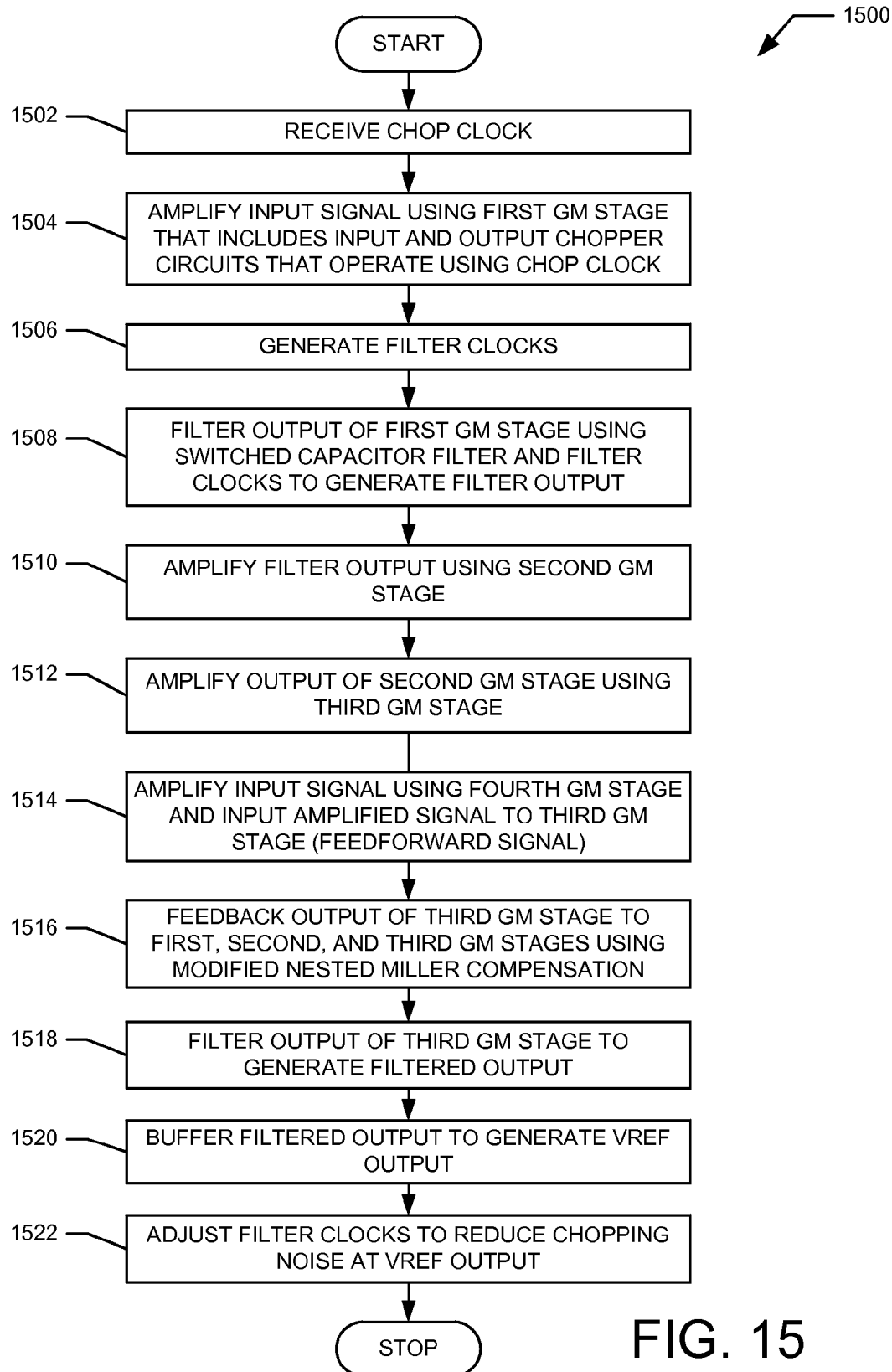


FIG. 15

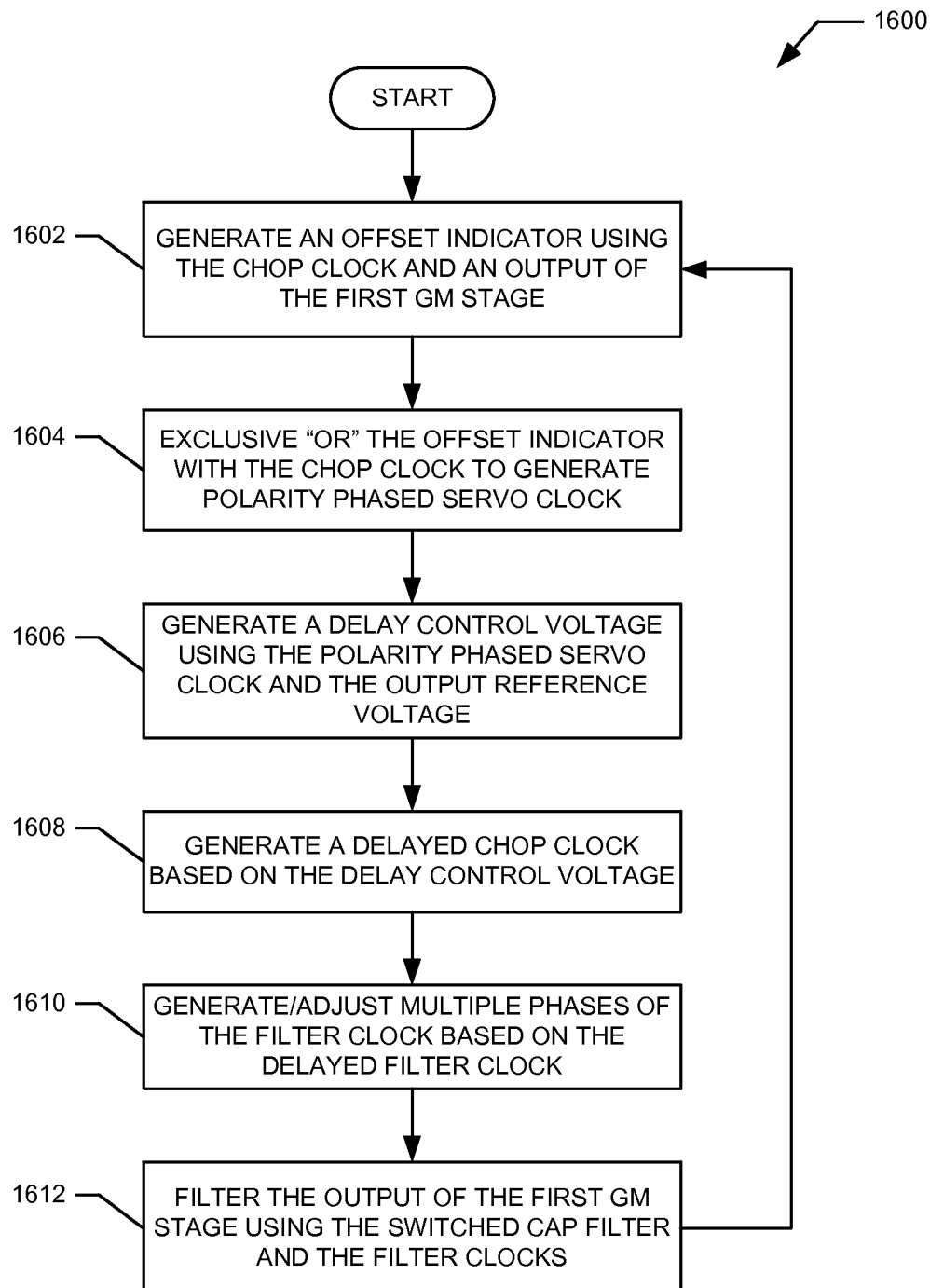


FIG. 16

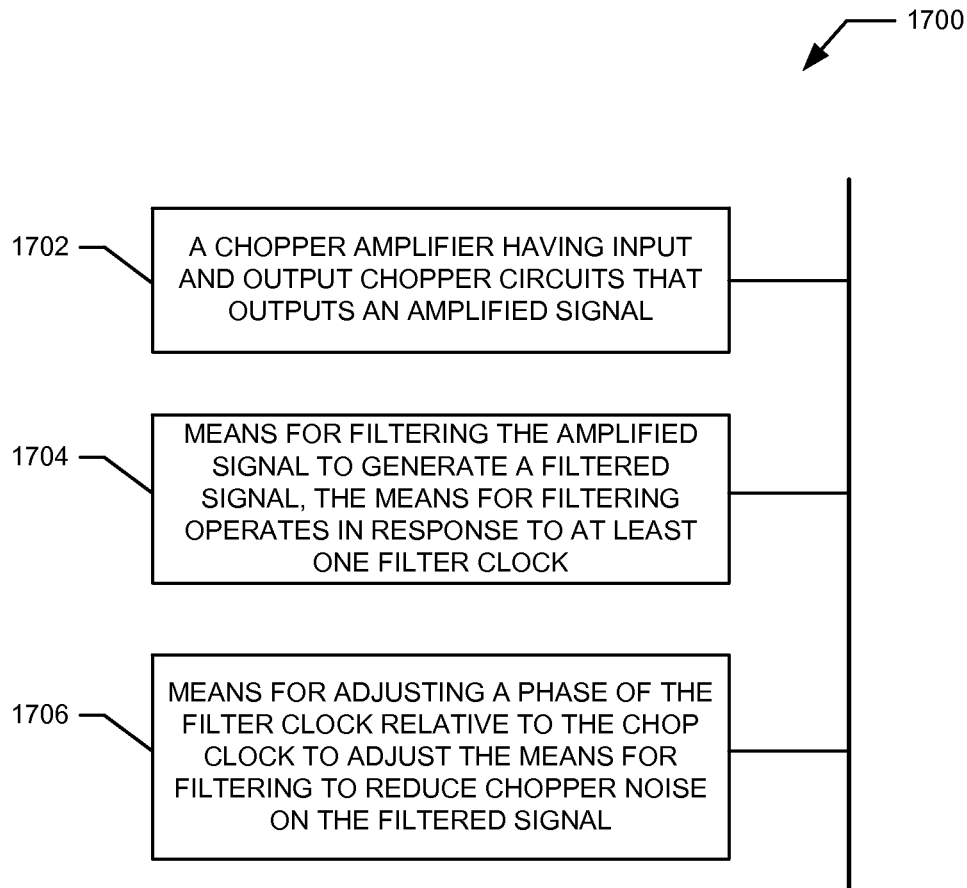


FIG. 17

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# CHOPPER STABILIZED AMPLIFIER WITH SYNCHRONOUS SWITCHED CAPACITOR NOISE FILTERING

## TECHNICAL FIELD

The present disclosure relates generally to chopper stabilized voltage amplifiers and more particularly to chopper stabilized amplifiers with synchronous filtering.

## BACKGROUND INFORMATION

Solid state reference voltage generators include an amplifier that is used to generate a Direct Current (DC) reference voltage. During operation, chopping an input stage of the amplifier can cause an offset of the amplifier to appear as a square or triangle wave signal at the chopping frequency at the amplifier output. In addition, voltage spikes at the chopper clock edges can propagate through to the output due to amplifier internal feedthrough and through the feedback network from the input to the output. When the amplifier is being used to produce a reference DC voltage, the square wave signal and voltage spike noise are undesirable. A more robust solution is desired.

## SUMMARY

A chopper stabilized amplifier with synchronous switched capacitor noise filtering is disclosed. In an exemplary embodiment, the novel amplifier reduces noise at the chopper frequency at the voltage output of the amplifier. A servo loop is included to continuously monitor the voltage output for noise at the chopper frequency. The servo loop continuously adjusts the delay of the capture time of a switched capacitor filter to optimally reduce the output noise at the chopper frequency. The amplifier also employs an ultra-wideband voltage follower stage at the amplifier output to absorb voltage spike noise at the output.

The foregoing is a summary and thus contains, by necessity, simplifications, generalizations and omissions of detail; consequently is it appreciated that the summary is illustrative only. Still other methods, and structures and details are set forth in the detailed description below. This summary does not purport to define the invention. The invention is defined by the claims.

## BRIEF DESCRIPTION OF THE DRAWINGS

The accompanying drawings, where like numerals indicate like components, illustrate embodiments of the invention.

FIG. 1 is a diagram of a digital to analog converter that includes an exemplary embodiment of a chopper stabilized amplifier with synchronous switched capacitor noise filtering.

FIG. 2 is an exemplary detailed embodiment of a bandgap voltage generator shown in FIG. 1 that includes the chopper stabilized amplifier with synchronous switched capacitor noise filtering.

FIG. 3 is an exemplary detailed embodiment of the chopper stabilized amplifier with synchronous switched capacitor noise filtering shown in FIG. 2.

FIG. 4 shows an exemplary detailed embodiment of a chopper amplifier shown in FIG. 3.

FIG. 5 shows an exemplary detailed embodiment of a switched capacitor filter shown in FIG. 3.

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FIG. 6 shows an exemplary embodiment of a filter timing adjuster shown in FIG. 3.

FIG. 7 shows an exemplary detailed embodiment of a non-overlap clock generator shown in FIG. 6.

FIG. 8 is an exemplary detailed embodiment of an offset sign detector shown in FIG. 6.

FIG. 9 is an exemplary detailed embodiment of a chopper noise null servo shown in FIG. 6.

FIG. 10 is an exemplary detailed embodiment of a voltage controlled delay shown in FIG. 6.

FIG. 11 shows timing diagrams that illustrate the operation of the offset sign detector shown in FIG. 6.

FIG. 12 shows timing diagrams that illustrate the operation of the chopper noise null servo shown in FIG. 6.

FIG. 13 shows timing diagrams that illustrate how a chop clock is delayed to generate the filter clocks that reduce chopper noise on an output voltage signal.

FIG. 14 shows a graph that illustrates the overall operation of an exemplary embodiment of a chopper stabilized amplifier with synchronous switched capacitor noise filtering to adjust the timing of the filter clocks relative to the chop clock to reduce chopper noise on an output voltage signal.

FIG. 15 shows a method for operating an exemplary embodiment of a chopper stabilized amplifier with synchronous switched capacitor noise filtering to reduce chopper noise on an output signal.

FIG. 16 shows an exemplary method for operating a filter timing adjuster in a chopper stabilized amplifier with synchronous switched capacitor noise filtering to reduce chopper noise on an output signal.

FIG. 17 is an exemplary embodiment of a chopper stabilized amplifier apparatus that generates and output with reduced chopper noise.

## DETAILED DESCRIPTION

FIG. 1 is a diagram of a digital to analog converter (DAC) system **100** that includes an exemplary embodiment of a chopper stabilized amplifier with synchronous switched capacitor noise filtering **102**. The system **100** includes DAC **104** and bandgap voltage generator (BG) **106**. During operation, the BG **106** generates a reference voltage signal (VREF) that is input to the DAC **104**. The DAC **104** uses the VREF signal to convert a digital code **108** to a corresponding output voltage **110**.

In an exemplary embodiment, the BG **106** includes the chopper stabilized amplifier with synchronous switched capacitor noise filtering **102**. For clarity, the chopper stabilized amplifier with synchronous switched capacitor noise filtering **102** is also referred to herein as "CSA" **102**. In an exemplary embodiment, the CSA **102** generates the VREF signal having less chopper noise when compared to conventional chopper amplifiers. This reduced noise level results in improved performance of the system **100**. More detailed descriptions of the design and operation of the CSA **102** are provided below.

FIG. 2 is an exemplary detailed embodiment of the bandgap voltage generator **106** shown in FIG. 1. As shown in FIG. 2, the BG **106** includes the CSA **102**.

The BG **106** includes transistors **206**, **208**, resistors **210**, **212**, and **214**, and capacitor **216**. In the exemplary embodiment shown in FIG. 2, the transistor **208** comprises 14 parallel transistors, each matched to transistor **206**.

The BG **106** generates a signal VBEH **202** at an emitter terminal of the transistor **206**, and also generates a signal VBEL **204** at an emitter terminal of the transistor **208**. The

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transistors **206** and **208** have base and collector terminals connected to a signal ground. The emitter terminal of the transistor **206** is also connected to a first terminal of the resistor **210**. A second terminal of the resistor **210** is connected to an output terminal **220**. The emitter terminal of the transistor **208** is also connected to a first terminal of the resistor **214** and a second terminal of the resistor **214** is connected to a first terminal of the resistor **212** and a first terminal of the capacitor **216**. Second terminals of the resistor **212** and capacitor **216** are connected to the output terminal **220**.

The voltage  $V_{BEH}$  **202** is applied to the positive input of CSA **102** as a  $V_{IN+}$  input. Because  $R_{F1}=R_{F2}$  and because the voltage between the + and - inputs to CSA **102** is very small, the  $V_{REF}$  signal that appears at node **220** will cause identical voltages across  $R_{F1}$  and  $R_{F2}$  and thus identical currents to flow through  $R_{F1}$  and  $R_{F2}$ . Because the current density is 14 times lower at  $Q_2$  than at  $Q_1$ ,  $V_{BEL}$  will be a lower voltage than  $V_{BEH}$  and the voltage difference  $DV_{BE}=(V_{BEH}-V_{BEL})=(kT/q)(\ln(14))$  will be dropped across resistor  $R_C$  **214**. Here  $k$  is Boltzmann's constant,  $q$  is the charge of an electron,  $T$  is temperature in degrees Kelvin and the factor of 14 is the ratio of the current densities at  $Q_1$  to  $Q_2$ .  $DV_{BE}$  increases as temperature increases.  $V_{BEH}$  decreases as temperature increases, approximately given by  $V_{BEH}=V_{GO}-(kT/q)\ln(I_0/I_{Q1})$  where  $V_{GO}$  is the bandgap voltage for silicon,  $I_0$  is a processing constant describing  $Q_1$  and  $I_{Q1}$  is the current flowing through the emitter of  $Q_1$ . When the ratio  $R_{F1}/R_C$  is chosen to generate the  $V_{REF}$  signal at node **220**, the voltage increase with temperature at node **220** due to  $DV_{BE}$  nearly exactly cancels the voltage decrease due to  $V_{BEH}$  **202** and the circuit implements a bandgap dc voltage reference.

The CSA **102** also receives a chop clock (FCHOP) **218** that is used to drive internal chopping circuits. The CSA **102** also includes switched capacitor filtering and other synchronization circuitry not shown in FIG. 2 but shown in greater detail in FIG. 3. During operation, the CSA **102** operates to generate the  $V_{REF}$  signal at node **220** exhibiting less chopper noise than conventional chopper stabilized amplifiers.

FIG. 3 is an exemplary detailed embodiment of the chopper stabilized amplifier with synchronous switched capacitor noise filtering **102** shown in FIG. 1 and FIG. 2. The CSA **102** comprises filter timing adjuster **304**, GM1 amplifier plus chop circuit **306**, switched capacitor filter **308**, GM2 amplifier **310**, GM3 amplifier **312**, GM4 amplifier **318**, filter stage **314**, buffer **316**, and modified nested miller compensation circuit **320**.

In an exemplary embodiment, a chop clock **322** is input to the GM1 amplifier plus chop circuit **306** and the filter timing adjuster **304**. In an exemplary embodiment, the chop clock **322** is a square wave having a frequency of approximately 400 KHz. The GM1 amplifier plus chop circuit **306** receives input voltages ( $V_{IN+}$ ,  $V_{IN-}$ ) and generates output voltages ( $V_{1+}$ ) **324** and ( $V_{1-}$ ) **326** that are input to the switched capacitor filter **308**. In an exemplary embodiment, the amplified signals  $V_{1+}$  **324** and  $V_{1-}$  **326** may include an amount of noise (e.g., chopper noise) as a result of the operation of the GM1 amplifier plus chop circuit **306**.

The filter **308** filters the input signals it receives to generate filtered output signals ( $V_{1F+}$ ) **328** and ( $V_{1F-}$ ) **330**. The filter **308** performs a filtering function utilizing filter clocks (PH1) **342** and (PH2) **344** received from the filter timing adjuster **304**. Due to variation of semiconductor processing steps, component mismatch and/or change in operating conditions, the delay from the chop clock to the

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chopper-caused noise at  $V_{1+}$ ,  $V_{1-}$  will vary. In an exemplary embodiment, the filter timing adjuster **304** operates to adjust the delay of the PH1 **342** and PH2 **344** clock signals to reduce the amount of chopper noise that appears at the output of the switched capacitor filter **308**.

The outputs (**328**, **330**) of the switch capacitor filter **308** are input to the GM2 amplifier **310** to generate outputs ( $V_{2+}$ ) **332** and ( $V_{2-}$ ) **334**. The outputs  $V_{2+}$  **332** and  $V_{2-}$  **334** from the GM2 amplifier **310** are input to the GM3 amplifier **312**. The input voltages ( $V_{IN+}$ ,  $V_{IN-}$ ) are also input to the GM4 amplifier **318** to generate outputs  $V_{4+}$  **346** and  $V_{4-}$  **348** that are also input to the GM3 amplifier **312**. The GM3 amplifier **312** amplifies the signals at its inputs to generate output  $V_3$  **336**. The amplifier GM4 provides a feed-forward path for the input signals that controls the output during transients. Amplifiers GM4 and GM2 are designed with very low open-loop gain. Amplifier GM1 is designed for very high open-loop gain. The amount of chopper noise suppression is proportional to the difference in the open-loop gains of these amplifiers, e.g., [noise suppression= $(A_{ol\ GM1})-(A_{ol\ GM4})$ ]. Amplifier GM3 combines the filtered and feed-forward signals and may be of only moderate open-loop gain because of the gain ( $A_{ol\ GM4}$ ) in front of it.

The output  $V_3$  **336** is input to a modified nested miller compensation circuit **320** that comprises capacitors **350**, **352**, **354**, **356**, and **360**. At a compensation input, a first terminal of the capacitor **354** is connected to the output  $V_3$  **336** and at a first compensation output, a second terminal of the capacitor **354** is connected to the inverting input terminal of the GM3 amplifier **312**. At the compensation input, a first terminal of the capacitor **352** is connected to the output  $V_3$  **336** and at a second compensation output, a second terminal of the capacitor **352** is connected to the non-inverting input terminal of the GM2 amplifier **310**. At the compensation input, a first terminal of the capacitor **350** is connected to the output  $V_3$  **336**, and at a third compensation output, a second terminal of the capacitor **350** is connected to the output  $V_{1+}$  **324** of the GM1 amplifier plus chop circuit **306**. A first terminal of the capacitor **356** is connected to the non-inverting input terminal of the GM2 amplifier **312** and a second terminal of the capacitor **356** is connected to a signal ground. The capacitor **360** is connected between the non-inverting output of the GM1 amplifier **306** and the signal ground.

The output  $V_3$  **336** of the GM3 amplifier **312** is input to the filter **314** to generate filtered output **338**. In an exemplary embodiment, the filter **314** is a simple RC filter. In other embodiments, the filter **314** is a more complex filter. The filtered output **338** is input to a non-inverting input terminal of the buffer amplifier **316** and buffered to generate the  $V_{REF}$  signal **358**. In an exemplary embodiment, the buffer amplifier **316** comprises a super wideband amplifier that absorbs noise glitches that may occur at the edges of the chop clock.

The filter timing adjuster **304** receives the  $V_{REF}$  signal **358**, the chop clock **322**, and the  $V_{1+}$  output **324** of the GM1 amplifier plus chop circuit **306**. The filter timing adjuster **304** utilizes the signals it receives to generate the filter clocks PH1 **342** and PH2 **344**.

During operation the filter timing adjuster **304** adjusts the phases of the filter clocks PH1 **342** and PH2 **344** with respect to the chop clock **322** to reduce or eliminate the amount of chopper noise that appears at the output of the filter **314**, and correspondingly, on the  $V_{REF}$  signal **358**.

FIG. 4 shows an exemplary embodiment of the GM1 amplifier plus chop circuit **306** shown in FIG. 3. In an

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exemplary embodiment the GM1 amplifier plus chopper circuit 306 includes an input chopper circuit 420 and an output chopper circuit 422. The input chopper circuit 420 comprises switches 402, 404, 406, and 408. The output chopper circuit 422 comprises switches 412, 414, 416, and 418. In an exemplary embodiment, the switches of the chopper circuits are implemented as transmission gates. The switches 402, 408, 412, and 418 are controlled by a first phase (CP1) of the chop clock 322. The switches 404, 406, 414, and 416 are controlled by a second phase (CP2) of the chop clock 322. For example, when the first phase (CP1) of the chop clock 322 is in a high voltage stage (logic high), the switches 402, 408, 412, and 418 are closed, and when the first phase (CP1) of the chop clock 322 is in a low voltage stage (logic low), the switches 402, 408, 412, and 418 are opened. The switches 404, 406, 414, and 416 are controlled by a second phase (CP2) of the chop clock 322 in a similar fashion.

The chop clock 322 is received at the GM1 amplifier plus chopper circuit 306 and is separated into the phase 1 chop clock (CP1) and the phase 2 chop clock (CP2) utilizing a non-overlapping clock generator (NOCG) 420. An exemplary embodiment and detailed description of the NOCG 420 is provided with reference to FIG. 7. The negative input voltage (VIN-) is received at the input of switches 402 and 404, and the positive input voltage (VIN+) is received at the input of switches 406 and 408. During operation, the switches 402, 404, 406, and 408 open and close in response to the CP1 and CP2 clocks. The output of switch 402 and the output of switch 406 are coupled together at a first input of GM1 amplifier 410. The output of switch 404 and the output of switch 408 are coupled together at a second input of the GM1 amplifier 410. The GM1 amplifier 410 operates to amplify the signals at its input terminals to generate amplified signals at its output terminals. In an exemplary embodiment, the GM1 amplifier 410 comprises a two-stage folded cascade amplifier.

A first output (inverted) of GM1 stage 410 is coupled to inputs of switch 412 and 414. A second output of GM1 stage 410 is coupled at the inputs of switch 416 and 418. During operation, the switches 412, 414, 416, and 418 open and close in response to the CP1 and CP2 clocks. The outputs of switches 412 and 416 are coupled together at the first device output to provide the V1+ 324 signal. The outputs of switches 414 and 418 are couple together at a second device output to provide the V1- 326 signal.

FIG. 5 shows an exemplary detailed embodiment of the switched capacitor filter 308 shown in FIG. 3. The switch capacitor filter 308 comprises switches 502, 504, 506, and 508. The switch capacitor filter 308 also comprises switches 510, 512, 514, and 516. In an exemplary embodiment, the switches of the filter 308 are implemented as transmission gates. The switch capacitor filter 308 receives the first phase (PH1) 342 and the second phase (PH2) 344 of the filter clock output from the filter timing adjuster 304. Each of the switches 502-516 opens and closes in response to the filter clock phase that it receives.

The filter 308 receives the voltage signals (V1+) 324 and (V1-) 326 from the output of the GM1 amplifier plus chopper circuit 306. The V1+ signal 324 is input to the switches 502 and 504. The V1- signal 326 is input to the switches 506 and 508.

The output of switch 502 is input to the switch 510 using signal line 518. The output of switch 504 is input to switch 512 using signal line 520. The output of switch 506 is input to switch 514 using signal line 522. The output 508 is input to switch 506 using signal line 524.

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A first capacitor 526 is connected between the signal line 518 and signal ground. A second capacitor 528 is connected between the signal line 520 and the signal ground. A third capacitor 530 is connected between the signal line 522 and the signal ground. A fourth capacitor 532 is connected between the signal line 524 and the signal ground. In an exemplary embodiment, the capacitors shown in FIG. 5 have capacitance values of approximately 4.3 Pico farads.

The outputs of switches 510, 512, 514, and 516 are coupled to first and second outputs of the switch capacitor filter 308. The first output of switch capacitor filter 308 provides the (V2+) 328 voltage output. A second output of switch capacitor 308 provides the (V2-) 330 voltage output. Capacitor 534 is connected between the V2+ 328 and V2- 330 outputs.

During the operation the input voltages 324 and 326 are switch through the filter 308 by the switches 502-516 to charge and discharge the capacitors 536, 528, 530, and 532. The switches are controlled by a first phase (PH1) 342 and a second phase (PH2) 344 of a filter clock. As a result, the output voltages (V2+) 328 and (V2-) 330 are generated and represent filtered versions of the input voltages. In an exemplary embodiment, the filter clocks PH1 and PH2 are phase adjusted with respect to the chop clock 322 by the filter timing adjuster 304 to reduce or eliminate the amount of chopper noise that appears on the output signals 328 and 330.

FIG. 6 shows an exemplary embodiment of the filter timing adjuster 304 shown in FIG. 3. In an exemplary embodiment the filter timing adjuster 304 comprises offset sign detector 602, XOR 604, chopper noise null servo 606, voltage controlled delay 610, and non-overlap clock generator 612.

In an exemplary embodiment, the offset sign detector 602 receives the chop clock 322 and the (V1+) 324 output. The offset sign detector 602 operates to generate a polarity indicator 620 that is input to XOR device 604. In an exemplary embodiment, the offset between the inputs of the chopped transimpedance amplifier GM1 306 causes triangular waves at the chop clock frequency at amplifier outputs V1+ and V1-. Positive polarity dc offsets at VIN+ to VIN- causes triangular waves that ramp upward during PH1 and back down during PH2. Negative polarity offset at VIN+ to VIN- causes triangular waves that ramp down during PH1 and up during PH2. The difference between the voltage at V1+ at the end of PH1 minus the voltage at the end of PH2 will be positive for positive dc offset at VIN+ to VIN-, and the difference will be negative for negative dc offset at VIN+ to VIN-. In an exemplary embodiment the peak to peak amplitude of the triangle wave is approximately 2.5 times the dc offset amplitude.

Thus the triangle wave at V1+ and a chop clock, which locally generates a copy of PH1 and PH2 (denoted PA1 and PA2) can be used to determine the polarity of the dc offset.

The XOR device 604 also receives the chop clock 322 as a second input. The output of the XOR device 604 is a polarity phased servo clock 622 that is an inverted chop clock for one detected polarity of the output of the offset sign detector 602 and a non-inverted chop clock for the other polarity of the output of the offset sign detector 602. In an exemplary embodiment, the XOR gate receives the chop clock 322 and the polarity indicator 620 and outputs the polarity phased servo clock 622 that is adjusted in phase by 0 or 180 degrees to reduce chopper noise for offsets of that polarity.

The chopper noise null servo 606 receives the polarity phased servo clock 622 and the VREF 358 signals and

generates a delay control voltage **616** that is input to the voltage control delay **610**. In an exemplary embodiment, the chopper noise null servo **606** generates the delay control voltage **616** such that the filter clocks (PH1, PH2) are phase adjusted to reduce chopper noise on the VREF **358** signal.

The voltage controlled delay **610** receives the chop clock **322** and the delay control voltage **616** from the chopper noise null servo **606**. The voltage controlled delay **610** delays the chop clock **322** in response to the delay control voltage **616** to generate a delayed chop clock **618**.

The non-overlap clock generator **612** receives the delayed chop clock **618** from the voltage control delay **610** and generates the first phase PH1 **342** and second phase PH2 **344** of the filter clock. The first and second filter clock phases PH1 and PH2 are used to adjust the switch timing of the switch capacitor filter **308** to eliminate or reduce chopper noise on the VREF **358** signal.

FIG. 7 shows an exemplary embodiment of the non-overlap clock generator **612** shown in FIG. 6. In an exemplary embodiment the non-overlap clock generator **612** comprises NAND gates **702** and **704**. The clock generator **612** also comprises inverters **706**, **708**, and **710**.

The delayed filter clock **618** is received at a first input of NAND gate **702** and a first input of inverter **706**. The output of the inverter **706** is input to a first input of NAND gate **704**. An output of NAND gate **702** is input to a second input of the NAND gate **704**. The output of NAND gate **704** is input to a second input of NAND gate **702**.

The output of NAND gate **702** is input to inverter **710** and the output of inverter **710** forms the first phase (PH1) **342** of the filter clock. The output of NAND gate **704** is input to inverter **708** and the output of inverter **708** forms a second phase (PH2) **344** of the filter clock. In an exemplary embodiment, the non-overlap clock generator **612** operates to provide a non-overlap (NO) region at the transitions of the output clock phases. The NO region is in the range of 500 picoseconds to 2-3 nanoseconds or more.

Thus, in various exemplary embodiments the non-overlap clock generator **612** operates to receive the delayed filter clock input **618** and generate multiple non-overlapping clock phases for use by the switch capacitor filter **308**.

It should also be noted that the structure of the non-overlapping clock generator **612** is also suitable for use as other non-overlapping clock generators described herein. For example, the non-overlapping clock generator **420** shown in FIG. 4, the non-overlapping clock generator **882** shown in FIG. 8, and the non-overlapping clock generator **964** shown in FIG. 9 can all be implemented using the same circuit structure as the non-overlapping clock generator **612** shown in FIG. 7.

FIG. 8 is an exemplary detailed embodiment of the offset sign detector **602** shown in FIG. 6. In an exemplary embodiment the offset sign detector **602** comprises a first AC coupled gain stage **802**, a second AC coupled gain stage **804**, a switched capacitor differentiator **806**, and an output stage **808**. The offset sign detector **602** also comprises a bias amplifier **880** and a non-overlapping clock generator **882**. The bias amplifier **880** receives a bias input signal at its non-inverting input. The bias input signal is created by the resistors **884** and **886** that are coupled between VDD and the signal ground. In an exemplary embodiment, the resistors **884** and **886** have identical resistor values, however, in other embodiments, the resistors **884** and **886** have different resistance values. The amplifier **880** generates a first common mode signal (VCM1) at its output and this first common mode signal is input to other sections of the offset sign detector **602** as described below.

In an exemplary embodiment, the non-overlapping clock generator **882** comprises circuitry similar to the non-overlapping clock generator **612** shown in FIG. 7. The NOCG **882** receives the chop clock **322** and generates non-overlapping phases of the chop clock designated (PA1) and (PA2), which are input to other sections of the offset sign detector **602** as described below.

During operation, the offset sign detector **602** receives the V1+ signal **324** and the VCM1 signal and determines the polarity indicator **620** that indicates how the filter clocks (PH1) **342** and (PH2) **344** are to be phase adjusted relative to the chop clock **322** to eliminate or reduce chopper noise on the VREF signal **358**.

The first AC coupled gain stage **802** comprises amplifier **810** that has a non-inverting input coupled to receive the VCM1 signal. The voltage V1+ **324** is input to a first terminal of resistor **812**. A second terminal of resistor **812** is connected to a first terminal of capacitor **814**. A second terminal of capacitor **814** is connected to an inverting input terminal of the amplifier **810**.

The inverting input terminal of the amplifier **810** is also connected to a first terminal of resistor **816** and a first terminal of capacitor **818**. An output terminal of the amplifier **810** is connected to a second terminal of the resistor **816** and a second terminal of the capacitor **818**.

A clamping circuit is formed by NMOS transistor **826** and PMOS transistor **822**. The output terminal of the amplifier **810** is connected to source terminals of the transistors **822** and **826**. The inverting input of the amplifier **810** is connected to drain and gate terminals of the transistor **822** and **826**. The clamping circuit formed by the transistors **822** and **826** operates to clamp the output of the amplifier **810**. In an exemplary embodiment, if the output of amplifier **810** is positive it is clamped to approximately the VGS of the PMOS transistor **822**. If the output of the amplifier **810** is negative, it is clamped to approximately the VGS of NMOS transistor **826**.

In an exemplary embodiment, the first AC coupled gain stage **802** provides a gain of approximately twenty (20) at the chop clock frequency and is AC coupled to remove DC offset.

In an exemplary embodiment the second AC coupled gain stage **804** comprises amplifier **828**. The output voltage of the amplifier **810** is connected to a first terminal of capacitor **830**. A second terminal of capacitor **830** is connected to a first terminal of resistor **832** and a non-inverting input terminal of the amplifier **828**. A switch **834** has first terminal that is connected to the non-inverting input of amplifier **828**. A second terminal of the resistor **832** is connected to a second terminal of the switch **834** and also receives the VCM1 signal. The switch **834** also includes a third terminal to receive a power up (PUP) signal that closes the switch when the system power reaches a desired operating level.

The inverting terminal of the amplifier **828** is connected to a first terminal of resistor **836**, a first terminal of resistor **838**, and a first terminal of capacitor **814**. A second terminal of the resistor **836** is connected to receive the VCM1 signal. An output terminal of the amplifier **828** is connected to a second terminal of the resistor **838** and a second terminal of the capacitor **840**.

A clamping circuit comprises NMOS transistor **848** and PMOS transistor **844**. The output terminal of the amplifier **828** is connected the source terminals of the transistors **848** and **844**. The inverting input terminal of the amplifier **828** is connected to drain and gate terminals of the transistors **848** and **844**. The clamping circuit formed by the transistors **848** and **844** operates to clamp the output of the amplifier **828**.

In an exemplary embodiment, if the output of amplifier **828** is positive it is clamped to approximately the VGS of the PMOS transistor **844**. If the output of the amplifier **828** is negative, it is clamped to approximately the VGS of NMOS transistor **848**.

In an exemplary embodiment, the second AC coupled gain stage **804** provides additional gain of approximately thirty-two (32) at the chop clock frequency.

In an exemplary embodiment, the switched capacitor differentiator **806** comprises an amplifier **850**. The output terminal of the amplifier **828** is connected to a first terminal of capacitor **852**. A second terminal of capacitor **852** is connected to a first terminal of switch **856**, a first terminal of switch **858**, and a first terminal of capacitor **854**. The second terminal of the switch **856** is connected to receive the VCM1 signal. A second terminal of capacitor **854** is connected to an inverting terminal of the amplifier **850** and a first terminal of the switch **860**. A non-inverting terminal of the amplifier **850** is connected to receive the VCM1 signal.

A second terminal of the switch **858** is connected to a second terminal of the switch **860**. The second terminal of the switch **860** is connected to a first terminal of the capacitor **862** and a first terminal of the switch **864**. The switch **858** has a third terminal connected to receive the PA2 clock, the switch **856** has a third terminal connected to receive the PA1 clock, and the switch **860** has a third terminal connected to receive the PA1 clock.

An output of the amplifier **850** is connected to a second terminal of capacitor **862** and a second terminal of the switch **864**. The switch **864** also has a third terminal connected to receive the PUP signal, which when active closes the switch **864**.

A clamping circuit is formed by NMOS transistor **872** and PMOS transistor **868**. The output terminal of the amplifier **850** is connected to source terminals of the transistors **872** and **868**. The inverting input of the amplifier **850** is connected to drain and gate terminals of the transistors **872** and **868**. The clamping circuit formed by the transistors **872** and **868** operates to clamp the output of the amplifier **850**. In an exemplary embodiment, if the output of amplifier **850** is positive it is clamped to approximately the VGS of the PMOS transistor **868**. If the output of the amplifier **850** is negative, it is clamped to approximately the VGS of NMOS transistor **872**.

In an exemplary embodiment, the switched capacitor differentiator **806** takes a sample on phase PA1 and a sample on phase PA2 and subtracts the difference while providing a gain of approximately three (3) and removing offset from the input signal utilizing correlated double sampling offset removal.

In an exemplary embodiment, the output stage **808** comprises low offset clocked comparator **874**, D-type flip-flop **876**, and D-type flip-flop **878**. The output terminal of the amplifier **850** is connected to a non-inverting terminal of the comparator **874**. An inverting input terminal of the comparator **874** is connected to receive the VCM1 signal. The comparator **874** also receives the PA2 clock, which operates to clock the comparator output signal.

An output terminal of the comparator **874** is connected to a D input of flip-flop **876**, and a Q output of flip-flop **876** is connected to a D input of flip-flop **878**. The PA1 clock is input to the clock terminal of flip-flop **876**. The flip-flop **878** receives the chop clock **822** at its clock terminal. The flip-flops **876** and **878** also receive the PUP signal at their reset inputs. The PUP signal operates to allow the flip-flops **876** and **878** to operate when power is established. The output of the flip-flop **878** is the polarity indicator **620**.

During operation, the comparator **874** latches the difference signal output from the switched capacitor differentiator **806** using the PA2 clock. After settling, the flip-flop **876** latches in response to the PA1 clock. The output of the flip-flop **876** is then resynched with the chop clock **322** to generate polarity indicator **620**.

FIG. 9 shows an exemplary detailed embodiment of the chopper noise null servo **606** shown in FIG. 6. In an exemplary embodiment the chopper noise null servo **606** comprises an AC coupled gain stage **902**, a switch capacitor differentiator **904**, and a switch capacitor integrator **906**. The chopper noise null servo **606** also includes bias amplifier **962** and NOCG **964**.

The bias amplifier **962** receives a bias input signal at its non-inverting input. The bias input signal is created by the resistors **966** and **968** that are coupled between VDD and the signal ground. In an exemplary embodiment, the resistors **966** and **968** have identical resistance values, however, in other embodiments, the resistors **966** and **968** have different resistance values. The amplifier **962** generates a second common mode signal (VCM2) at its output and this second common mode signal is input to other sections of the chopper noise null servo **606** as described below.

In an exemplary embodiment, the non-overlapping clock generator **964** comprises circuitry similar to the non-overlapping clock generator **612** shown in FIG. 7. The NOCG **964** receives the polarity phased servo clock **622** and generates non-overlapping phases of the polarity phased servo clock **622** that are designated (PB1) and (PB2), which are input to other sections of the chopper noise null servo **606** as described below.

In an exemplary embodiment, the chopper noise null servo **606** operates to receive the VREF **358** and VCM2 signals and generates the delay control signal (VCNTL) **616**, which controls phase adjustment of the PH1 and PH2 filter clocks relative the chop clock **322** to eliminate or reduce chopper noise on the VREF **358** signal.

In an exemplary embodiment, the AC coupled gain stage **902** receives the VREF **358** and VCM2 signals and provides an AC gain of approximately (x250) around the chop clock frequency. The switched capacitor differentiator **904** takes the difference between the voltage at the output of amplifier **908** at phase PB1 and the voltage at phase PB2 and provides an amplified difference signal at the output of amplifier **922**. The switched capacitor integrator **906** performs a cumulative sum of those differences. Initially, the differences may be large but will step closer and closer to a settling point.

In an exemplary embodiment, the AC coupled gain stage **902** comprises amplifier **908** that receives the VCM2 signal at a non-inverting input terminal. The VREF signal **358** is received at a first terminal of resistor **910**. A second terminal of the resistor **910** is connected to a first terminal of capacitor **912**. A second terminal of capacitor **912** is connected to an inverting input of the amplifier **908**.

The inverting input terminal of the amplifier **908** is connected to a first terminal of resistor **914** and a first terminal of capacitor **920**. A second terminal of the resistor **914** is connected to a first terminal of resistor **918** and a first terminal of resistor **916**. A second terminal of resistor **918** is connected to receive the VCM2 signal. An output of the amplifier **908** is connected to a second terminal of resistor **916** and a second terminal of capacitor **920**. During operation, the AC coupled gain stage **902** provides a gain of at least two hundred (200) at the frequency of the chop clock **322**.

In an exemplary embodiment, the switched capacitor differentiator **904** comprises amplifier **922** having a non-

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inverting input terminal connected to receive the VCM2 signal. The output of the amplifier 908 is input to a first terminal of capacitor 924. A second terminal of capacitor 924 is connected to a first terminal of switch 928, a first terminal of switch 930, and a first terminal of capacitor 926.

A second terminal of switch 928 is connected to receive the VCM2 signal and a third terminal of the switch 928 is connected to receive the PB1 clock. A second terminal of capacitor 926 is connected to a first terminal of switch 932 and an inverting input terminal of the amplifier 922. A second terminal of the switch 930 is connected to a second terminal of switch 932, a first terminal of switch 936, and a first terminal of capacitor 934.

Amplifier 922 has an output terminal that is connected to a second terminal of capacitor 934, a second terminal of switch 936. The switch 928 and the switch 932 are opened and closed in response to the PB1 clock signal. Switch 930 is open and closed in response to the PB2 clock signal. The switch 936 is open and closed in response to the PUP signal.

A clamping circuit is formed by PNP bipolar transistor 938 and PNP bipolar transistor 940. The output terminal of the amplifier 922 is connected to an emitter terminal of transistor 940 and a base terminal of transistor 938. Collector terminals of the transistors 938 and 940 are connected to signal ground. An emitter terminal of transistor 938 and a base terminal of transistor 940 are connected to the inverting input of the amplifier 922. The clamping circuit formed by the transistors 938 and 940 operates to clamp the output of the amplifier 922. In an exemplary embodiment, if the output of the amplifier 922 is positive, the output is clamped to approximately the VBE of transistor 940. If the output of amplifier 922 is negative, the output is clamped to approximately the VBE of transistor 938.

In an exemplary embodiment, the switched capacitor differentiator 904 provides a gain of approximately fourteen (14), and correlated double sampling offset removal. The switched cap differentiator 904 computes  $V(\text{PH1}) - V(\text{PH2})$  which serves as a synchronous detector for voltage changes between PH1 and PH2, indicative of chopper noise.

In an exemplary embodiment the switch capacitor integrator 906 comprises amplifier 942 that has a non-inverting input terminal connected to receive the VCM2 signal. An output of amplifier 922 is connected to a first terminal of switch 944. A second terminal of the switch 944 is connected to a first terminal of switch 948, and a first terminal of capacitor 946. A second terminal of capacitor 946 is connected to a first terminal of switch 950, a first terminal of switch 952 and a first terminal of capacitor 960.

A second terminal of capacitor 960 is connected to a first terminal of switch 954 and an inverting input of amplifier 942. A second terminal of switch 950 is connected to receive the VCM2 signal. A second terminal of switch 954 is connected to a second terminal of switch 952, a first terminal of switch 958, and a first terminal of capacitor 956. And output terminal of amplifier 942 is connected to a second terminal of capacitor 956 and a second terminal of capacitor 958. The output terminal of the amplifier 942 also outputs the VCNTL signal 616.

The switch 948 and the switch 952 are opened and closed in response to the PB1 signal. The switch 950, the switch 954, and the switch 944 are open and closed in response to the PB2 signal. The switch 958 is normally open and thereafter closed when the PUP signal is active.

In an exemplary embodiment, the switched capacitor integrator 906 provides a gain of approximately  $(\frac{1}{2})$  and correlated double sampling offset removal. Integrator 906 generates a cumulative sum of the output of the differentiator

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904, which is then output as VCNTL 616. When the chopper noise is large the integrator will generate large steps in VCNTL 616 to quickly begin to reduce the chopper noise. As the chopper noise is reduced the integrator generates smaller and smaller steps in VCNTL 616. If the loop should start up with too much VCNTL 616 the differentiator will generate an output opposite to the usual sign and the integrator will generate steps in the negative direction until the correct VCNTL 616 has been reached and the loop settles.

In an exemplary embodiment, the switched capacitor differentiator 904 and integrator 906 will generate a VCNTL voltage 616 that increases or decreases when there is a positive step voltage at VREF voltage during chopper noise null servo clock phase CP1 depending upon the polarity indicator 620. The correct polarity will cause the differentiator 904 and integrator 906 to generate a VCNTL 616 signal that exhibits negative feedback, which is desired. The wrong polarity will cause differentiator 904 and integrator 906 to generate a VCNTL 616 that exhibits positive feedback within the chopper servo null loop and the chopper servo null loop may increase, not decrease, chopper noise at the VREF output.

FIG. 10 shows an exemplary detailed embodiment of the voltage controlled delay 610 shown in FIG. 6. In an exemplary embodiment, the chop clock 322 is input into a gate terminal of transistor 1016 and a gate terminal of transistor 1014, which form inverter buffer 1028. The delay control signal (VCNTL) 616 is received and input to a first terminal of resistor 1002. A second terminal of the resistor 1002 is connected to a drain terminal of transistor 1004, a gate terminal of transistor 1004, a gate terminal of transistor 1006, and a gate terminal of transistor 1018. The transistors 1004, 1006, and 1008 form current mirror 1024. Source terminals of transistors 1004, 1006, and 1018 are connected to a signal ground. A drain terminal of transistor 1006 is connected to drain and gate terminals of transistor 1010. A source terminal of transistor 1010 is connected to drain and gate terminals of transistor 1008. A source terminal of transistor 1008 is connected to a supply voltage (VDD).

The gate terminal transistor 1008 is connected to a gate terminal of transistor 1012. A source terminal of the transistor 1012 is connected to the supply voltage and a drain terminal of the transistor 1012 is connected to a source terminal of transistor 1014. The transistors 1008 and 1012 form current mirror 1026. A drain terminal of transistor 1014 is connected to a drain terminal of transistor 1016. A source terminal of transistor 1016 is connected to a drain terminal of transistor 1018.

A drain terminal of transistor 1014 is connected to gate terminals of transistors 1020 and 1022, which form inverter buffer 1030. Capacitor 1032 has a first terminal connected to the drain terminal of transistor 1014 and a second terminal connected to the signal ground. A source terminal of transistor 1020 is connected to the supply voltage and a source terminal of transistor 1022 is connected to the signal ground. Drain terminals of the transistors 1020 and 1022 are connected together at an output node 1034 and the delayed filter clock signal 618 is output from the output node 1034. In an exemplary embodiment, the resistor 1002 has a resistance of approximately 85 KOhm and the capacitor 1032 has a capacitance value of approximately 0.12 Pico farads.

During operation, the VCNTL voltage signal 616 is converted to a current by resistor 1002 and this current signal is input to the current mirror 1024. The current flowing through transistor 1006 is mirrored by the current flowing through transistors 1004 and 1018. The current

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flowing through transistor **1006** also flows through transistor **1008** and this current is mirrored by the current flowing through the transistor **1012**.

The chop clock **322** is input to the inverter buffer **1028**, which acts like a switch to allow current to flow through transistor **1014** to the capacitor **1032** or from the capacitor through transistor **1016**. Thus, the capacitor **1032** effectuates a delay. A triangle waveform is generated at the node that is common to the drains of the transistors **1014** and **1016**. The inverter buffer **1030** then converts this triangle wave to the delayed filter clock **618**.

FIG. **11** shows timing diagrams that illustrate the operation of the offset sign detector **602** shown in FIG. **6**. During operation, the  $V_{1+}$  **324** signal shown at **1102** is received at the first AC coupled gain stage **802**. The signal is amplified as discussed above and the output of the amplifier **810** is shown at **1104**. This signal is then input to the second AC coupled gain stage **804** and the output of the amplifier **828** is shown at **1106**. The output of the second AC coupled gain stage **804** is input to the switched capacitor differentiator **806** and the output of amplifier **850** is shown at **1108**. This output represents the polarity indicator **620**. It should be noted that the power up process holds the output of amplifier **850** low for approximately the first 22 microseconds.

FIG. **12** shows timing diagrams that illustrate the operation of the chopper noise null servo **606** shown in FIG. **6**. During operation, the  $V_{REF}$  **358** signal shown at **1202** is input to the first AC coupled gain stage **902**. The output of the first AC coupled gain stage is provided at the output of the amplifier **908**, which is shown at **1204**. The output of the amplifier **908** is input to the switched capacitor differentiator **904**, which generates an amplified difference between the first and second clock phases (P2A, P2B). This difference is provided at the output of the amplifier **922** and shown at **1206**. The output of the amplifier **922** is input to the switched capacitor integrator **906**. The switch capacitor integrator **906** generates the  $V_{CNTL}$  signal **616** at the output of the amplifier **942**, which is shown at **1208**.

FIG. **13** shows timing diagrams that illustrate how the chop clock **322** is adjustably delayed to generate the filter clocks that reduce chopper noise on the  $V_{REF}$  signal. In a first graph **1302**, the chop clock **322** is shown at **1304**. In the early stages of operation (e.g., 20 microseconds after startup) the chop clock **322** is delayed to generate a delayed chop clock **618** (shown at **1306**) that is used to generate the filter clocks (PH1, PH2). As can be seen in graph **1302** the initial delay **1314** is approximately 21 nanoseconds.

In a second graph **1308**, the chop clock **322** is shown at **1310**. After approximately 100 microseconds of operation, the chop clock **322** is adjustably delayed by the generation of the  $V_{CNTL}$  signal to generate a delayed chop clock **618** (shown at **1312**) that is used to generate the filter clocks (PH1, PH2). As can be seen in graph **1308** the delay **1316** has changed to approximately 19 nanoseconds. The system will thereafter continuously adjust the delay to reduce chopper noise on the output of the filter **308** and thus on the  $V_{REF}$  signal.

FIG. **14** shows a graph that illustrates the overall operation of the CSA **102** to adjust the timing of the filter clocks relative to the chop clock to reduce chopper noise on the  $V_{REF}$  signal. As illustrated in FIG. **14**, the positive output of the GM1 stage **306** is a triangle waveform **1402** that is input to the switched capacitor filter **308**. The switched capacitor filter **308** operates in response to the first (PH1) and second (PH2) filter clocks, which operate in the regions **1404** and **1406**, respectively. As illustrated in FIG. **14**, the filtering does not align with the triangle wave's passage

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through ground, a small error voltage results, which is illustrated at **1412** and **1418**. During operation of the CSA **102**, the filter clocks are adjustably delayed such that the filtering during phase 1 and phase 2 aligns with the triangle wave's passage through ground. For example, the filter clocks are adjusted by the time delays shown at **1408** and **1410**. The adjustment to the delay of the filter clock phases results in a reduction in the delay time error and shrinks the corresponding voltage error, during both filter phases. For example, the error voltages **1414** and **1416** represent a square wave error voltage that appears on the filter output during both phases, and which is gradually reduced as the filter clocks (PH1, PH2) are time adjusted. As a result, the chopper noise on the  $V_{REF}$  signal is reduced.

FIG. **15** shows a method **1500** for operating an exemplary embodiment of a chopper stabilized amplifier with synchronous switched capacitor noise filtering to reduce chopper noise on the output signal. In an exemplary embodiment, the method is suitable for use with the CSA **102** shown in FIG. **3**.

At block **1502**, a chop clock is received. In an exemplary embodiment the chop clock **322** is received by the CSA **102** shown in FIG. **2** and FIG. **3**.

At block **1504**, an input signal is amplified using a first GM stage that includes input and output chopper circuits that operate using the received chop clock. For example, the chopped GM stage **306** amplifies the input voltages ( $V_{IN+}$ ,  $V_{IN-}$ ) to generate an output signal ( $V_{1+}$ ) **324** and ( $V_{1-}$ ) **326** using input (**420**) and output (**422**) chopping circuits and the GM stage **410**.

At block **1506**, switched capacitor filter clocks are generated. In an exemplary embodiment the non-overlapping clock generator **612** generates the switched capacitor filter clocks (PH1, PH2) based on a delayed version of the chop clock **322**.

At block **1508**, the output of the chopped GM stage is filtered by a switched capacitor filter using the filter clocks generated at block **1506**. In an exemplary embodiment the output ( $V_{1+}$ ,  $V_{1-}$ ) of the chopped GM stage **306** is filtered using the switch capacitor filter **308** that operates utilizing the filter clocks (PH1, PH2) to generate a filtered output ( $V_{1F+}$ ,  $V_{1F-}$ ).

At block **1510**, the outputs of the switched capacitor filter are amplified using a second (GM) stage. In an exemplary embodiment, the outputs ( $V_{1F+}$ ,  $V_{1F-}$ ) of the switch capacitor filter **308** are amplified using the GM2 stage **310** to generate outputs ( $V_{2+}$ ) **332** and ( $V_{2-}$ ) **334**.

At block **1512**, the outputs of the GM2 stage are amplified using a 3rd (GM) stage. In an exemplary embodiment, the outputs ( $V_{2+}$ ,  $V_{2-}$ ) from the GM2 stage **310** are input to the GM3 stage **315** to generate an amplified ( $V_3$ ) output signal **336**.

At block **1514**, the input voltages are amplified by a fourth (GM) stage to generate amplified voltages that are input to the GM3 stage. In an exemplary embodiment, the input voltages ( $V_{IN+}$ ,  $V_{IN-}$ ) are input and amplified by the GM4 stage **318** to generate amplified voltages ( $V_{4+}$ ) **346** and ( $V_{4-}$ ) **348** that are input to the GM3 stage **312**. The signals  $V_{4+}$  and  $V_{4-}$  represent feedforward signals to the GM3 stage.

At block **1516**, the output of the GM3 stage is fed back to previous GM stages utilizing a modified nested miller compensation circuit. In an exemplary embodiment, the output ( $V_3$ ) **336** of the GM3 stage **312** is input to the modified nested miller compensation circuit **320**. The capacitor **354** provides a feedback path to the input of the GM3 stage **312**. The capacitor **352** forms a feedback path to the input of the

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GM2 stage **310**. The capacitor **350** forms a feedback path to the input of the filter **308**. The capacitor **356** couples the non-inverting input terminal of the GM3 stage **312** to the signal ground.

At block **1518**, the output of the GM3 stage is filtered using a low pass filter. In an exemplary embodiment, the (V3) output **336** of the GM3 stage **312** is filtered using the single pole filter **314** to generate the filter output **338**.

At block **1520**, the output of the single pole filter is buffered to generate the voltage reference signal (VREF). In an exemplary embodiment, the output **338** of the filter **314** is buffered by the wideband buffer **316** to generate the reference voltage (VREF) signal **358**.

At block **1522**, the filter clocks that are input to the switch capacitor filter are adjusted to reduce chopper noise on the VREF signal. In an exemplary embodiment, the filter timing adjuster **304** adjusts the phases of the filter clocks (PH1 and (PH2) with respect to the chop clock **322** to reduce the level of chopping noise that appears in the VREF signal. A more detailed description of the operations performed at block **1522** is provided in the description of the method **1600** shown in FIG. **16**.

FIG. **16** shows an exemplary method **1600** for operating a filter timing adjuster in a chopper stabilized amplifier with synchronous switched capacitor noise filtering to reduce chopper noise on the output signal. In an exemplary embodiment, the method is suitable for use with the filter timing adjuster **304** shown in FIG. **6**. In an exemplary embodiment, the method **1600** is suitable for use at block **1522** of the method **1500**.

At block **1602**, an offset indicator is generated using the chop clock and an output of the first GM stage. In an exemplary embodiment, the offset sign detector **602** operates to receive the chop clock **322** and the V1+ **324** output of the GM1 stage **306** and generates the polarity indicator **620** as described above.

At block **1604**, the polarity indicator is gated with the chop clock using an XOR gate to generate a polarity phased servo clock. In an exemplary embodiment, the polarity indicator **620** and the chop clock **322** are input to XOR **604**. The XOR **604** generates the polarity phased servo clock **622** at its output port.

At block **1606**, a delay control voltage is generated using the polarity phase servo clock and the output reference voltage VREF. In an exemplary embodiment, the chopper noise null servo **606** received the polarity phased servo clock **622** and the reference voltage VREF **358** and generates the delay control voltage **616**.

At block **1608**, a delayed chop clock is generated based on the delay control voltage. In an exemplary embodiment, the voltage controlled delay **610** receives the chop clock **322** and the delay control voltage **616** and generates the delayed chop clock **618**.

At block **1610**, multiple phases of the filter clock are generated (or adjusted) using the delayed chop clock. In an exemplary embodiment, the non-overlap clock generator **612** received the delayed chop clock **618** and generates multiple phases (PH1 **342**, PH2 **344**) of the filter clock.

At block **1612**, the output of the GM stage **306** is filtered using the switch capacitor filter that is operated utilizing the filter clocks. In an exemplary embodiment switched capacitor filter **308** receives the output of the first GM stage **306** and filters this output utilizing the filter clocks PH1 **340** and PH2 **344**. By utilizing the filter clocks the amount of chopper noise that appears at the output voltage reference VREF is eliminated or reduced. In an exemplary embodiment, the phase/timing of the filter clocks PH1 **340** and PH2 **344** are

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continually adjusted by operation of the method **1600** such that the reduced chopper noise level on the VREF signal **358** is maintained. In an exemplary embodiment, the method continues at block **1602**.

FIG. **17** is an exemplary embodiment of a chopper stabilized amplifier apparatus **1700** that generates an output with reduced chopper noise. For example, the apparatus **1700** is suitable for use as the CSA **102** shown in FIG. **3**.

The apparatus **1700** comprises a chopper amplifier having an input that receives an input signal and an output that outputs an amplified signal. The chopper amplifier includes an input chopping circuit coupled to the input and an output chopping circuit coupled to the output, the input and output chopping circuits operate in response to a chop clock. In an exemplary embodiment, the chopper amplifier comprises the GM1 amplifier plus chop circuit **306**.

The apparatus **1700** also comprises a first means (**1704**) for means for filtering the amplified signal to generate a filtered signal. The means for filtering operates in response to at least one filter clock. In an exemplary embodiment, the means for filtering comprises a switched capacitor filter operated with filter clocks. For example, the means for filtering **1704** comprises the switched capacitor filter **308**.

The apparatus **1700** also comprises a second means (**1706**) for adjusting a phase of the filter clock relative to the chop clock to adjust the means for filtering to reduce chopper noise on the filtered signal. In an exemplary embodiment, the means **1706** comprises the filter timing adjuster **304**.

Although certain specific embodiments are described above for instructional purposes, the teachings of this patent document have general applicability and are not limited to the specific embodiments described above. Accordingly, various modifications, adaptations, and combinations of various features of the described embodiments can be practiced without departing from the scope of the invention as set forth in the claims.

What is claimed is:

1. An apparatus comprising:

a chopper amplifier having an input that receives an input signal and an output that outputs an amplified signal, the chopper amplifier includes an input chopping circuit coupled to the input and an output chopping circuit coupled to the output, the input and output chopping circuits operate in response to a chop clock;

a switched capacitor filter having an input that receives the amplified signal and an output that outputs a filtered signal, the switched capacitor filter operates in response to at least one filter clock; and

a filter timing adjuster having an input to receive a reference voltage and the chop clock, and an output to output the at least one filter clock, the filter timing adjuster adjusts a phase of the at least one filter clock with respect to the chop clock to reduce chopper noise on the reference voltage.

2. The apparatus of claim 1, wherein the filter timing adjuster generates a first filter clock (PH1) and a second filter clock (PH2), and wherein the filter timing adjuster adjusts a phase of the PH1 clock and a phase of the PH2 clock with respect to the chop clock to reduce chopper noise on the filtered signal.

3. The apparatus of claim 2, wherein the filter timing adjuster comprises:

an offset sign detector that receives the chop clock, and the amplified signal, and generates a polarity indicator; and

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an exclusive-OR (XOR) gate that receives the chop clock and the polarity indicator and outputs a polarity phased servo clock adjusted in phase by 0 or 180 degrees to reduce chopper noise for offsets of that polarity.

4. The apparatus of claim 3, wherein the filter timing adjuster comprises:

- a chopper noise null servo that receives the polarity phased servo clock and the reference voltage and generates a delay control signal;
- a voltage controlled delay that receives the chop clock and the delay control signal and generates a delayed chop clock; and
- a non-overlap clock generator that receives the delayed chop clock and generates the PH1 filter clock and the PH2 filter clock.

5. The apparatus of claim 1, wherein the chopper amplifier comprises a first GM stage.

6. The apparatus of claim 5, further comprising:

- a second GM stage that receives a filter output signal from the switched capacitor filter and generates a second amplified signal;
- a third GM stage that receives the second amplified signal and a fourth amplified signal and generates a third amplified signal; and
- a buffer stage that receives a filtered version of the third amplified signal and generates the reference voltage.

7. The apparatus of claim 6, further comprising:

- a fourth GM stage receives the input voltage and generates the fourth amplified signal.

8. The apparatus of claim 7, further comprising:

- a modified nested miller compensation circuit having a compensation input that receives the third amplified signal reference voltage and having a first compensation output that outputs a first feedback signal that is input to the switched capacitor filter, a second compensation output that outputs a second feedback signal that is input to the second GM stage, and a third compensation output that outputs a third feedback signal that is input to the third GM stage.

9. The apparatus of claim 8, wherein the modified nested miller compensation circuit comprises:

- a first capacitor connected between the compensation input and the first compensation output;
- a second capacitor connected between the compensation input and the second compensation output;
- a third capacitor connected between the compensation input and the third compensation output;
- a fourth capacitor connected between a non-inverting input of the third GM stage and a signal ground; and
- a fifth capacitor connected between an output of the first GM stage and the signal ground.

10. The apparatus of claim 1, wherein the apparatus forms a bandgap reference voltage generator.

11. An apparatus comprising:

- a chopper amplifier having an input that receives an input signal and an output that outputs an amplified signal, the chopper amplifier includes an input chopping circuit coupled to the input and an output chopping circuit coupled to the output, the input and output chopping circuits operate in response to a chop clock;

means for filtering the amplified signal to generate a filtered signal, the means for filtering operates in response to at least one filter clock; and

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means for adjusting a phase of the filter clock relative to the chop clock to adjust the means for filtering to reduce chopper noise on the filtered signal.

12. The apparatus of claim 11, wherein the means for adjusting generates a first filter clock (PH1) and a second filter clock (PH2), and wherein the means for adjusting adjusts a phase of the PH1 clock and a phase of the PH2 clock with respect to the chop clock to reduce chopper noise on the filtered signal.

13. The apparatus of claim 12, wherein the means for adjusting comprises:

- means for generating a polarity indicator from the chop clock and an amplified input signal; and
- means for exclusive-ORing (XOR) the chop clock and the polarity indicator to generate a polarity phased servo clock that is adjusted in phase by either 0 or 180 degrees to reduce chopper noise for offsets of that polarity.

14. The apparatus of claim 13, wherein the means for adjusting comprises:

- means for generating a delay control signal from the polarity phased servo clock and a reference voltage;
- means for generating a delayed chop clock from the chop clock and the delay control signal; and
- means for generating non-overlapping versions of the PH1 and PH2 filter clocks from the delayed chop clock.

15. A method comprising:

- amplifying an input signal using a chopper amplifier to generate an amplified signal, wherein the chopper amplifier includes an input chopping circuit coupled to receive the input signal and an output chopping circuit coupled to the output the amplified signal, the input and output chopping circuits operate in response to a chop clock;

- filtering the amplified signal to generate a filtered signal, wherein the filtering is performed using at least one filter clock; and

- adjusting a phase of the at least one filter clock relative to the chop clock to reduce chopper noise on a reference voltage.

16. The method of claim 15, wherein the operation of adjusting comprises generating a first filter clock (PH1) and a second filter clock (PH2), and adjusting a phase of the PH1 clock and a phase of the PH2 clock with respect to the chop clock to reduce chopper noise on the reference voltage.

17. The method of claim 16, wherein the operation of adjusting comprises:

- generating a polarity indicator from the chop clock and an amplified input signal; and
- exclusive-ORing (XOR) the chop clock and the polarity indicator to generate a polarity phased servo clock that is adjusted in phase by either 0 or 180 degrees to reduce chopper noise for offsets of that polarity.

18. The method of claim 17, wherein the operation of adjusting comprises generating a delay control signal from the polarity phased servo clock and the reference voltage.

19. The method of claim 18, wherein the operation of adjusting comprises generating a delayed chop clock from the chop clock and the delay control signal.

20. The method of claim 19, wherein the operation of adjusting comprises generating non-overlapping versions of the PH1 and PH2 filter clocks from the delayed chop clock.

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